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Some Diamagnetic Problems

ACCORDING to the dictionary diamagnetic means 'cross-magnetic'—applied to any substance, such as a rod of bismuth or glass, which, when suspended between the poles of a magnet, arranges itself across the line joining the poles' in contradistinction to paramagnetic; i.e., parallel to the line joining the poles. In the *Philosophical Magazine*, April 1855, however, there was published a letter from Professor Thomson (later Lord Kelvin) to Professor Tyndall in which he maintained that if a rod of bismuth is placed in a magnetic field obliquely across the lines of force, 'the resultant of the couples will tend to turn the whole mass in such a direction as to place the length of the bar along the lines of force.' He then said 'Some phenomena that have been observed are to be explained by the resultant of forces from places of stronger to places of weaker intensity in the field, others by the resultant of couples depending on crystalline structure, and others by the resultant of such forces and couples co-existing; and none observed depend at all on any other cause.' In 1852 in a report to the British Association at Belfast Thomson had said that 'what is known with certainty regarding magnetic induction and magnetic action generally, shows that an elongated mass, *in a uniform magnetic field*, tends to place its length parallel to the lines of force, whether its inductive capacity be ferromagnetic or diamagnetic, provided it be non-crystalline, because if ferromagnetic it becomes more, or if diamagnetic, less intensely magnetized, if placed in such a position, than if placed with its length across the lines of force.' The italics are not in the original, but that Thomson realized their importance is obvious from his further remark that 'All directional tendencies in bars of diamagnetic substance which have yet been, and probably all that can ever be discovered by experiment are thus due to the tendency of their ends or other movable parts, *from places of stronger towards places of weaker force*, in varied magnetic fields.' (These italics are in the original.) Five years before this, in 1847, when his address was St. Peter's College, Cambridge, Thomson wrote that 'such a needle has no tendency to arrange itself across the lines of magnetic force, but if it be very small compared with the dimensions and distance of the magnet, the direction it will assume, when allowed to turn freely, about its centre of gravity, will be that of the lines of force, whether the material of which it consists be diamagnetic or magnetic matter such as soft iron; but Faraday's result is due to the rapid decrease of magnetic intensity round the poles of the magnet, and to the length of the needle, which is considerable compared with the distance between the poles of the magnet; and is thus explained by the discoverer [Faraday] himself, "The cause of the pointing of the bar, or any oblong arrangement of the heavy glass is now evident. It is merely a result of the tendency of the particles to move outwards, or into the positions of weakest magnetic action."'

Although Brugmans had discovered in 1778 that a piece of bismuth repelled a magnet pole,

it was not until 1845 that Faraday showed its diamagnetic properties and invented the name.

And now, a century later, the fallacy still persists and has recently been the topic of letters in the American *Electrical Engineering*. Slepian raised the question in Sept. 1947 and gave the correct answer in Jan. 1948. In May Litman suggested an interesting experiment which is an exact converse of Faraday's experiment. By using pointed poles Faraday produced a field that was strong at the axis and fell off rapidly at a short distance from the axis. Litman suggests that if a magnet be made as shown in Fig. 1 and a paramagnetic or ferromagnetic rod be suspended at the centre, it should tend to set itself transverse to the field, especially if knobs be fitted to the ends of a thin rod. This would be due to the tendency of the ends of the rod to move into the stronger field, just as in Faraday's experiment the ends of the bismuth rod moved into the weaker field.

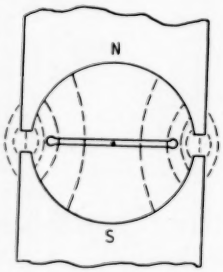

Fig 1.

If one consults text-books of fifty years ago one finds a very confused state of affairs. Some are just vague, some support the view held by Kelvin—S. P. Thompson, for example, says 'There is no reason to doubt that in a magnetic field of uniform strength a bar of bismuth would point along the lines of induction,'—while others teach exactly the opposite. Watson in his "Text-book of Physics" says that 'in a *uniform* field there is no measurable directive force . . . but the manner in which a diamagnetic cylinder will set itself in a very strong magnetic field can, however, be at once foreseen.' He then shows with the help of a diagram that 'diamagnetic bodies therefore tend to turn so that their longer axis is at right angles to the direction of the field.' Mascart and Joubert's "Electricity and Magnetism" (1883) says 'The coefficient of magnetization for diamagnetic bodies is so feeble that the reactions of the particles may be neglected and their effect escape all means of observation. For a diamagnetic needle, provided it is not crystallized, is in mobile equilibrium in a uniform magnetic field; in all experiments in which there seem to be phenomena of direction, the effect is due to the *magneto-crystalline* properties of the body.'

All this is now a thing of the past, and although many people still believe that a diamagnetic rod, if free to rotate, takes up a transverse position in a uniform field, this is disproved in all reliable text-books that give serious consideration to magnetic theory.

In the *Phil. Mag.* for January 1856, there is a letter from Thomson addressed to Tyndall giving his answer to a question propounded by the latter. The question was 'Supposing a cylinder of bismuth to be placed within a helix, and surrounded by an electric current of sufficient intensity; can you say, *with certainty*, what the action of either end of that cylinder would be on an external fragment of bismuth presented to it?' To which Thomson replied 'In answer, I say, that the fragment of bismuth will be *repelled* from either end of the bar provided the helix be infinitely long . . . The confidence amounts in my own mind to a feeling of *certainty*.' Now we have a feeling of certainty that Thomson was wrong. He says 'In every case in which the forces experienced by a little magnetized steel needle held with its axis reverse along the lines of force, and a fragment of bismuth substituted for it in the same locality of a magnetic field, have been compared, they have been found to agree . . . In no case has any discrepance from this law been observed . . . In every case in which any magnetic force at all can be observed on a fragment of bismuth, it is such as the steel magnet thus held experiences. Therefore I say it is in this case repulsion.' This may have satisfied Tyndall but we think that it was a wrong conclusion. Fig. 2(a) shows a paramagnetic rod and Fig. 2(b) a diamagnetic rod in the magnetic field of a long solenoid. If the rod were absolutely non-magnetic the field would not be distorted in any way and B in the rod would be $\mu_0 H_0$ where μ_0 is the permeability of space.

Fig. 2

With a paramagnetic rod the field tends to crowd into the rod and with a diamagnetic rod it tends to avoid it. In both cases $B = \mu_0 H + 4\pi\kappa H$. In the first case H in the rod is less than H_0 because of the demagnetizing effect of the poles but, notwithstanding this, B is greater than $\mu_0 H_0$ because the susceptibility κ is positive. In the second case H in the rod is greater than H_0 because of the magnetizing effect of the poles but, notwithstanding this, B is less than $\mu_0 H_0$ because the susceptibility κ is negative.

Now everyone agrees that in Fig. 2(a) a frag-

ment of iron placed at X would move to the right and a fragment of bismuth to the left, the bismuth tending to move to the weaker field. A magnetized needle held reversed at X would, as Thomson said, do the same as the bismuth. Although we talk of attraction and repulsion it is really a matter of moving to a stronger or weaker field. In Fig. 2(b) a fragment of iron placed at Y would move to the left and a fragment of bismuth to the right, because by moving to the right the bismuth would move into a weaker field. A magnetized needle held reversed at Y would similarly move to the right into the weaker field. Hence the bismuth rod would appear to attract the fragment of bismuth and not repel it as Thomson maintained. The small loop and arrow-head on the rod merely indicate the direction of the electronic currents within the material. One must not be surprised that in Fig. 2(b) flux enters the north pole and leaves the south pole. This flux is the resultant of that due to the diamagnetized rod and the much greater flux due to the helix. In Fig. 2(a) the magnetic induction just above and below the rod is weakened by the action of the poles, whereas in Fig. 2(b) it is strengthened. Anyone who dislikes poles can regard it as due to the action of the electronic currents within the material of the rod.

The same conclusion is reached by the following reasoning. If both the cylinder and the fragment are of iron, there is attraction, and if the iron fragment is replaced by a fragment of bismuth, there is repulsion. Everyone agrees on this. If now the iron cylinder is replaced by one of bismuth, its polarity is reversed. and the repulsion must surely be replaced by attraction. It is difficult to see how Thomson came to a contrary conclusion.

Probably the simplest way of showing why a bismuth rod in a uniform field sets itself along the direction of the field is to follow up what we said above about the effect of the poles in increasing the value of H in the rod. In a paramagnetic material the poles exert a demagnetizing force and therefore tend to weaken the field, expecially if the path through the material is short and the poles therefore close together. In a diamagnetic material under such circumstances the poles exert a magnetizing force. One must remember that we are dealing with materials that hardly affect the applied field, and that conditions are very different from those that obtain with a ferromagnetic

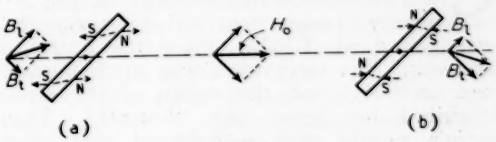

Fig. 3

rod. Fig. 3 shows (a) a paramagnetic and (b) a diamagnetic rod placed at 45° to the direction of the field H_0. In the latter the polarities are reversed, and, whereas in the former the longitudinal magnetization B_l is greater than the transverse B_t, the reverse is true in the diamagnetic case. The path of the magnetic induction through the materials and the corresponding poles will be as shown. In both cases the forces exerted on the poles by the field, which are indicated by the arrows, produce a clockwise torque tending to bring the rod into line with the applied field. This is practically the same as the method of explaining the phenomenon given by Thomson in his letter to Tyndall. In Fig. 3 we have greatly exaggerated the difference between B_l and B_t ; in the actual material the difference between them will be very small.

G. W. O. H.

TRANSMISSION-LINE IMPEDANCE MEASUREMENT

Balanced Lines at 100-1,000 Mc/s

By R. J. Lees, M.A., B.Sc.,* C. H. Westcott, Ph.D.*† and F. Kay, M.A.*‡

1. Introduction

CONSIDERABLE experience of aerial design and impedance measurement in the frequency range from 100 to 1,000 Mc/s was gained at T.R.E. during the course of the war. The early work was at frequencies near 200 Mc/s, on the design of the aerials described by Taylor and Westcott.[1] These aerials mostly used a balanced twin open-wire feeder of characteristic impedance $Z_0 = 330$ ohms, although for some receiving aerials a 75-ohm coaxial feeder was used. The earliest impedance measurements were made using the method of Cork and Pawsey,[2] but difficulties were experienced, particularly with twin feeders, for which the attainment of a truly balanced condition proved difficult. The alternative method involving the measurement of standing-wave patterns was therefore increasingly adopted, and the present paper describes the technique evolved for obtaining satisfactory results with balanced-feeder systems using this method, including the developments which made the method reliable at higher frequencies.

The alternative methods described by Essen[3] were probably developed somewhat before ours, but the two developments were essentially independent. It is interesting to note that we both experienced similar difficulties, in particular the presence of unbalanced components on balanced feeders. We believe that our methods are particularly suited to aerial design problems and other measurements where the standing-wave ratio is fairly close to unity, although for other purposes, such as cable-loss measurements we are willing to concede the superiority of Essen's methods. The self-consistency of the results which we have obtained gives us considerable confidence in the accuracy of our methods; the results given by Westcott and Goward[4] on the impedances of various dipoles in parabolae may be cited as an example.

Although it must be admitted that measurements on a balanced feeder are more difficult than unbalanced measurements, we have generally rejected the alternative of measuring a balanced impedance by using a balance-to-unbalance line transformer and an unbalanced measuring device. The reason for this is that discontinuity effects associated with the transformer may cause serious and incalculable departures from the theoretical transformation laws. This effect also occurs if the transformer is not correctly adjusted for the frequency used. Where for special reasons we have used such a transformer with an unbalanced measuring equipment, we have preferred the 'Pawsey Stub'[5] type shown in Fig. 1 (a) to the half-wave loop [or 'trombone'] Fig. 1 (b), or the slotted-coaxial type, Fig. 1 (c).

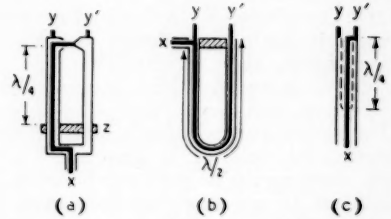

Fig. 1. *Balance-to-unbalance devices: (a) Pawsey stub; (b) half-wave trombone; (c) slotted coaxial stub; x = unbalanced feed; y, y' = balanced feed; z = sliding shorting bar.*

The Pawsey stub, if designed so that the outer conductors are symmetrical, gives perfect balance,§ although if the shorting-bar (z) is incorrectly adjusted a shunt reactance is placed across the load being measured; the adjustment is, however, readily made if the load is disconnected. The trombone is more difficult to construct without introducing asymmetrical feeder discontinuities, and if the loop is not exactly a half-wavelength long unbalanced components arise on the balanced feeder. It may

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§ Perfect balance is when the potentials V_1 and V_2, at corresponding points on the two lines are equal and opposite. In the general case we write $V_1 = U + B$ and $V_2 = U - B$, where U and B are the unbalanced and balanced components respectively. Either component may be analysed into harmonics (if these occur) in the usual way.

sometimes be of use when a 4 : 1 balance-to-unbalance impedance transformation is required; the Pawsey stub, of course, gives 1 : 1. Other arrangements, such as the slotted coaxial of Fig. 1 (c), while sometimes used in aerial design, are unsuitable for measurement purposes since the transformations they introduce are incalculable.

We therefore recommend making the measurements on a balanced system whenever practicable. The precautions necessary to ensure true balance and detect unbalanced components are described below. It is important to eliminate effects due to unbalanced components and harmonics since the aerial under test, though matched to its feeder for the wanted component, may present a very different impedance to other components, reflecting some of them strongly and giving rise to large standing waves. It is equally desirable to ensure that the measuring device is relatively insensitive to such unwanted components.

2. Early Work at 200 Mc/s

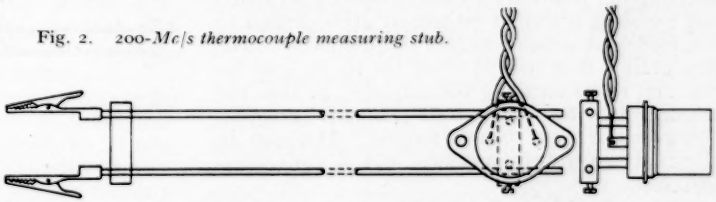
Our earliest work at frequencies of about 200 Mc/s was done using the Cork and Pawsey[2] method, in which the detuning and damping effect of the load when placed across a tuned LC circuit was compared with the effect of standard resistances and reactances. At this frequency, however, this equipment is approaching the limit of its usefulness for several reasons. In the first place it is difficult to make a truly symmetrical tuned circuit and to energize it in a properly balanced manner. Secondly, standard resistors, especially the higher values, no longer present the same impedance as they do at low frequencies. Further, as generally constructed, the equipment is most sensitive when measuring relatively high impedances, and can be used with a quarter-wave transformer to measure low impedances, but is least suited to measuring the medium impedances usually presented by aerials.

For reasons of convenience and accuracy, we therefore developed the standing-wave measurement method for balanced lines. In design work on aerials the standing-wave ratio‖ is usually fairly small (say not exceeding 2), which makes this method convenient. As in the case of normal coaxial-line measurements, the position of the standing wave on the feeder and its ratio depend uniquely on the load impedance (for a feeder of given Z_0), so that by measuring the ratio and the position of (say) the minimum of the standing-wave pattern the impedance of the load can be determined. The calculation is most readily made on a 'Smith chart' or circle diagram,[6] which is a geometrical representation of the standard transmission-line formula.

The aerials on which most of our measurements were made were fed by 330-ohm open-wire twin feeder consisting of 200-lb hard-drawn copper wires (0.112-in diameter) spaced 0.875 in between centres. Usually the feeders were tensioned, a minimum of Distrene spacers being used, and as a later refinement, these spacers were used in pairs $\lambda/4$ apart so that the reactance introduced by one of the pair was cancelled to the first order by that due to the other.

To measure the r.f. voltage across the feeder at any point, a clip-on stub arrangement was used (Fig. 2). The stub is $\lambda/4$ long and is short-circuited, so that it presents a substantially infinite impedance at the point of attachment and does not upset the voltage distribution on the feeder by drawing current. However, a considerable wattless current flows in the shorting-bar, given by $i = V/Z_0$, where V is the voltage across the points of attachment to the feeder and Z_0 the characteristic impedance of the stub. Due to the self-inductance of the shorting-bar there is a small voltage between the ends of this bar, and this is measured by connecting a vacuum thermocouple across it. The current drawn by the thermocouple heater is very small compared with the wattless current in the bar so that the load on the feeder remains negligible; the power absorbed by the thermocouple is naturally abstracted from the feeder system, but it constitutes only a negligible fraction of the power passing along the line to the load. An oscillator with a mean output of the order of 1 watt or more was necessary to energize the feeder system, and a sensitive thermocouple was used.

Fig. 2. *200-Mc/s thermocouple measuring stub.*



The thermocouple was connected to a millivoltmeter (2.4 mV full scale) for taking measurements; the reading of this meter was of course proportional to the power dissipated in the

‖ Throughout this paper, this ratio is defined as that of the voltage maximum to the voltage minimum (i.e., it is never less than unity).

thermocouple ; i.e., to the square of the voltage across the feeder. As the stub was moved along the feeder the standing-wave pattern could be measured. When plotted, a sine-wave curve should result§ [$V^2 \propto a + b \sin(\pi x/\lambda + c)$] from which the ratio and position of the standing wave can be determined. It is generally desirable to plot the complete pattern from measurements made at uniformly-spaced points along a line, rather than just to look for and measure the maximum and minimum, since the presence of harmonics may be suspected if the curve is not truly sinusoidal. No special precautions had to be taken to screen the leads to the thermocouple, although erratic results and severe hand-capacitance effects were obtained when using faulty thermocouples in which electrical continuity existed between the heater and couple elements. Essen experienced difficulties due to the same cause.

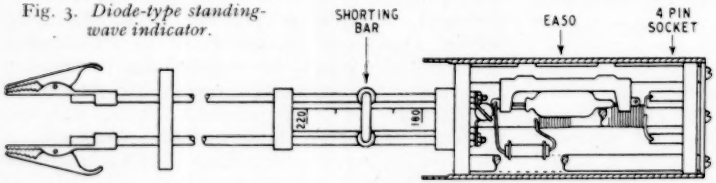


Fig. 3. *Diode-type standing-wave indicator.*

As will be seen from Fig. 2, the position of the shorting-bar is adjustable for working at different frequencies. The determination of the correct shorting-bar position for each frequency is carried out by using a second measuring stub clipped to the feeder at a point nearer the oscillator. The first stub is then moved along the feeder and the position of its shorting bar adjusted so that the reading of the other stub remains constant however the first stub is moved. This can be done at several frequencies and a calibration obtained.

Although this equipment is satisfactory for laboratory use, it possesses several disadvantages for tests on actual radar installations. For such work it is advantageous to be able to use the radar transmitter itself as the generator. The thermocouples are easily damaged by overload and also require the use of sensitive low-impedance galvanometers. For field use, therefore, a more robust equipment was designed making use of an EA50 type diode instead of a thermocouple. The mechanical arrangement of this is shown in Fig. 3, and the circuit in Fig. 4. The smoothing circuit following the diode rectifier is designed to work with the pulses given by the radar transmitter (3-5μsec duration, repetition rate 500 c/s). The choke stores energy during the pulse, providing a back-e.m.f., and this, together with the 200-ohm series resistance, limits the peak diode current and reduces the tendency to overload, as well as improving the linearity and rendering the sensitivity of the system less dependent on the diode-cathode temperature. The choke also serves to lengthen the current pulse through the diode and gives improved sensitivity compared with that which would be obtained with a series resistance giving an equivalent back-e.m.f. When used with a 6,000-ohm meter giving full-scale deflection for 15 volts, a very nearly linear calibration law (meter reading against r.f. voltage on the feeder) was obtained, although for accurate work the instrument was calibrated against a stub fitted with a thermocouple. A standing-wave pattern plotted with a diode measuring stub is not an exact sinusoid since the reading is proportional to V and not V^2. The minimum is therefore sharper than the maximum, which makes it preferable to measure the position of the former rather than the latter to obtain greater accuracy.

It will be seen from symmetry that these measuring stub arrangements are insensitive to unbalanced voltages on the feeder, the end of the stub where the diode or thermocouple is connected being a current node for the unbalanced components. The r.f. chokes fitted to the diode model help in this respect, reducing any sensitivity

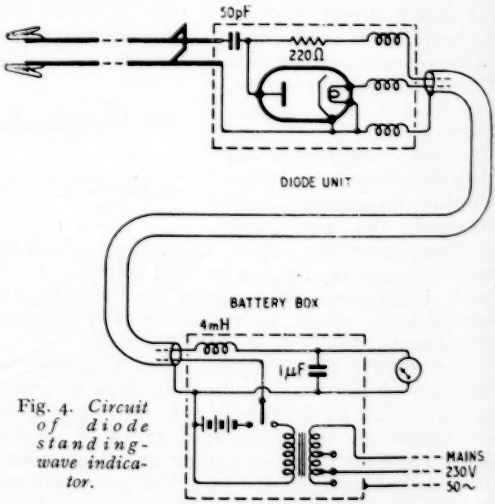


Fig. 4. *Circuit of diode standing-wave indicator.*

§ "Standing Waves and Impedance Circle Diagrams", by C. H. Westcott, Ph.D. To be published in *Wireless Engineer*.

due to the unbalanced voltage antinode. The trouble experienced with faulty thermocouples with a heater to thermocouple connection has already been mentioned; this fault might also be expected to lead to sensitivity to the unbalanced component. Although in theory this stub is sensitive to balanced harmonic components no trouble was experienced due to these in the 200-Mc/s measurements, since it was found to be fairly easy to design oscillators and coupling systems which eliminated harmonics. It is, of course, important when using the measuring stub to attach it so that it is perpendicular to the feeder, or when working near feeder junctions, to keep it well clear of the other feeders, to avoid mutual inductance effects between the feeder and the measuring stub. These would cause errors due to a sensitivity to the currents flowing along the line, which have a different standing-wave pattern from the voltages across it.

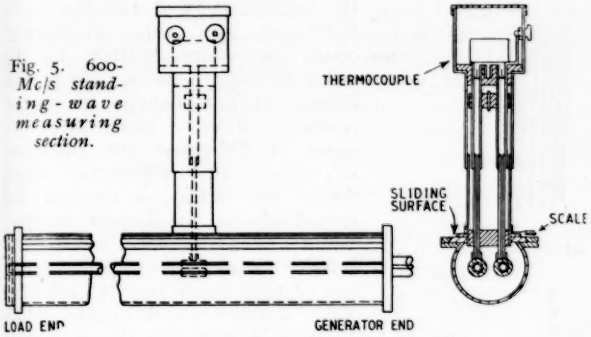


Fig. 5. 600-Mc/s standing-wave measuring section.

It should be mentioned that besides its use for measuring impedances, this stub finds application in measuring the power on feeders. It was, in fact, first evolved in connection with such an application. The power passing along a feeder is given by:

$$\text{Power} = \frac{V_{max} \times V_{min}}{Z_0}$$

so that relative power measurements only require a measure of the maximum and the minimum of the standing wave. This can be used for monitoring transmitter output, or for checking the power distribution to the separate sections of a complex array; e.g., in setting up a 'tapered' array for the reduction of side lobes (cf. Ratcliffe[5] and Taylor and Westcott[2]). The stub was also used in the design of the capacitance switch described by the latter authors and in other work on r.f. switching, for checking the matching through the switch and the switching efficiency.

3. Measurements in 500-600 Mc/s Band

The aerials used for frequencies in the 500-600 Mc/s band were generally fed with screened twin feeders. Polythene-dielectric cable was often used in production, but for experimental work a standard form of fabricated line was adopted. This consisted of two copper tube feeders spaced $\frac{3}{4}$in between centres surrounded by a screen of tube of outside diameter 2in, Distrene spacers being used as necessary. The diameter of the inner tubes was chosen according to the characteristic impedance required (usually 100-130 ohms). For making standing-wave measurements on this form of line, a special measuring section was designed which could be attached to, and form part of, the main feeder. It consists of a length of line of similar dimensions to this feeder, but with a slot cut in the outer to allow a measuring stub to project through and slide along. The slot is arranged symmetrically with respect to the inner conductors and a subsidiary experiment has shown that its effect on the characteristic impedance of the section is negligible. Contact between the ends of the stub and the inner conductors is made through smoothly sliding shoes, and the stub is screened within a tube fastened to a carriage which slides between guide-rails and carries the whole stub with it (see Fig. 5). The whole measuring stub is telescopic so that its length can be adjusted for working at different frequencies. In order to reduce the sensitivity of the stub to harmonics, the shorting-bar used at 200 Mc/s is replaced by a capacitance formed by two brass blocks insulated by a slip of mica. The spacing of these blocks determines the capacitive impedance which shunts the thermocouple, and thereby fixes the sensitivity of the measuring stub. Its value is usually such that a stub length of about 0.26λ is needed to give infinite impedance. The use of a capacitance instead of a shorting-bar helps to reduce the sensitivity of the stub to harmonics, since the impedance in parallel with the thermocouple is one the magnitude of which decreases with increasing frequency, in contrast with the behaviour of an inductive shorting-bar. The use of a capacitance also avoids a resonance for even harmonics; a simple shorted $\lambda/4$ stub would behave for these modes like a short-circuit across the line and most of the power available in them would be fed into the thermocouple. It is also important to choose the spacing of the capacitor blocks from the outer tube to avoid such a resonance for unbalanced modes.

To reduce further the unwanted modes, a line filter was used between the oscillator and the measuring section. This is shown in Fig. 6. The pick-up system in the oscillator compartment consists of a balanced tuned circuit, which is very helpful in rejecting harmonics. The output of this circuit is fed through two equal coaxial cables into the balancing unit. In this, two trombones are joined between the two lines at a distance of $\lambda/4$ apart.¶ These trombones bridges are adjusted to be $\lambda/2$ long, so that the voltages at their ends are equal and opposite. They act as a short-circuited quarter-wave stub to the balanced component, which therefore passes unhindered, but for the unbalanced component they behave as short-circuits, the two in succession being a very effective filter. These stubs also remove balanced even harmonics, but unbalanced even harmonics are passed by them; usually this did not lead to any trouble, but on occasions when it did, the unwanted modes were easily removed by adding separate $\lambda/4$ shorted stubs to each of the two coaxial feeders. A convenient point to attach such stubs is mid-way between the balancing trombones.

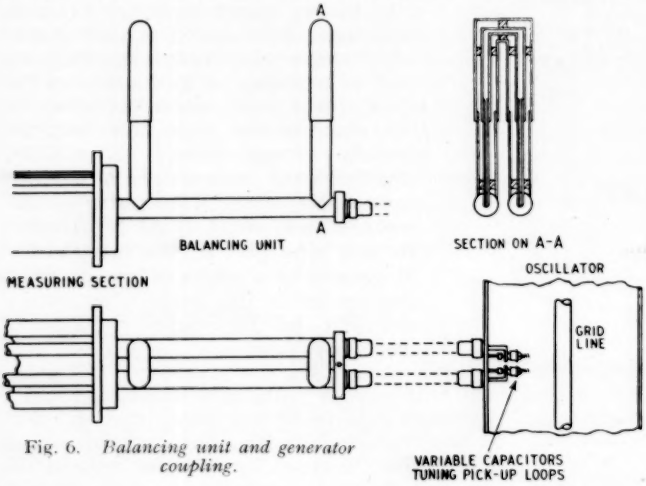


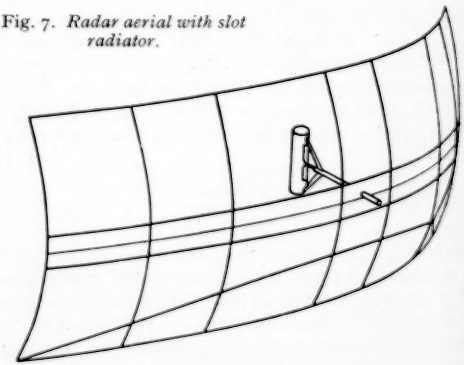
Fig. 6. *Balancing unit and generator coupling.*

The measuring stub and the filter elements were all made telescopic and could be adjusted for use at various frequencies. The tubes were all calibrated, the adjustment of the measuring stub length being made by substantially the same method as described for the 200 Mc/s equipment in Section 2. Prior to this, the oscillator frequency was adjusted using an accurate wavemeter, and the balancing stubs set approximately. The pick-up circuits were also tuned to give a maximum reading on the thermocouple; an h.t. voltage control for the oscillator was also provided so that the magnitude of the thermocouple meter reading could be adjusted to a convenient value. The balance of the system could be checked by attaching a further $\lambda/2$ trombone across the feeder at a suitable point; this trombone was like those¶ in the balancing filter, but the mid-point of its inner conductor could be short-circuited to its outer through a small hole in the latter. If any change in meter readings occurred when this was done, it indicated the presence of unbalanced components, since when the stub is shorted its impedance to unbalanced components changes from zero to infinity while its impedance to the wanted mode remains infinite. As a further check on the reliability of the results the standing-wave pattern was plotted—as at 200 Mc/s it should be sinusoidal.

4. Experimental Results

Some of the results obtained with this equipment have already been published,[4] and some further examples are given in this section. It was required to design a radar aerial giving horizontal polarization for any frequency in the range 500–600 Mc/s with a minimum of standing wave on the feeders, and giving a horizontal beam-width of 5–6° and a vertical beam-width of about 20°. The form of aerial used to meet this specification was a wire-netting reflector of aperture 26-ft wide and 7-ft 6-in high, consisting of two 13-ft wide rectangular sections of

Fig. 7. *Radar aerial with slot radiator.*



¶ A similar arrangement has recently been described by J. Tomiyasu (*Rev. Sci. Instrum.*, 1948, Vol. 19, p. 675). We were using this device early in 1943.

paraboloids of revolution whose corners lay in the focal plane. Each half of the aerial was fed with a single radiator near the focus of the reflector. It is necessary that such a radiator should radiate over almost 180° in the horizontal plane in order to illuminate the reflector properly, while in the vertical plane the radiation should be confined to a 110° sector in order to avoid loss of power past the edges of the reflector. A vertical half-wave dipole and reflecting plate approximately fulfilled these requirements but gave vertically polarized radiation. To reduce

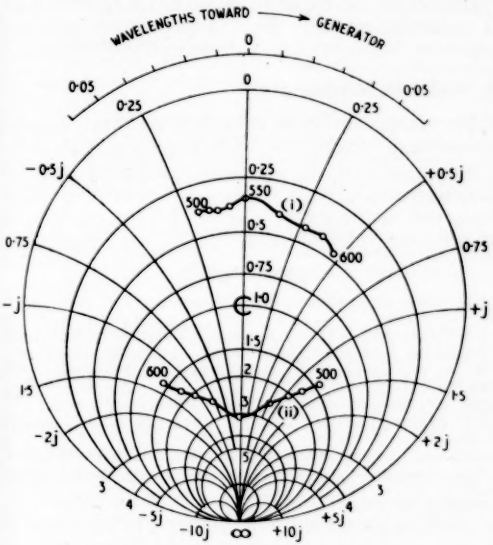


Fig. 8. *Variation of admittance of ideal slot with frequency, without paraboloid. Curve* (i) *admittance at slot; curve* (ii) *admittance at end of transformer; both normalized to* 260 *ohms. The frequencies of the* 500-, 550- *and* 600 *Mc/s points are marked; intermediate points at* 12.5 *Mc/s intervals. C = centre of diagram.*

fixed ground echoes and sea returns horizontal polarization was required, and therefore a vertical slot radiator,[7, 8] cut in a hollow metal cylinder was used. This gives nearly the same radiation pattern as a vertical dipole but is horizontally polarized.

The feeder to be used was a 125-ohm twin polythene cable, but, owing to the high impedance of a slot radiator, a quarter-wavelength of open-wire line ($Z_0 = 260$ ohms) was first attached to the slot. For experimental measurements this was connected to a 4-ft 6-in length of fabricated 140-ohm feeder, which was clamped to the netting reflector (one 13-ft section only) so that the distance of the slot from the apex of the reflector could be varied (cf. Fig. 7). The measuring section was then connected to the feeder at the back of the paraboloid. The cylinder used was 21.5-in long and 5.4-in diameter (λ and $\lambda/4$ respectively at 550 Mc/s), and a 2-in wide slot was found to be resonant at this frequency at a length of 13.4in (0.624λ). Its impedance was then about 720 ohms. Fig. 8

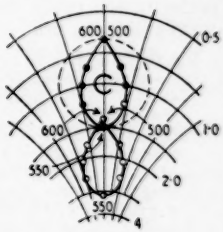


Fig. 9. *Frequency variation of admittance of ideal slot with compensating stub (no paraboloid); lower curve normalized to* 260 *ohms; upper curve normalized to* 125 *ohms. The* 500- *and* 600-*Mc/s points are marked; intermediate points are at* 12.5 *Mc/s intervals. The dotted circle represents a standing wave of* 1.5 : 1.

shows the type of variation of admittance (the reciprocal of impedance) with frequency obtained for a slot of this type in free space plotted on a circle diagram in terms of a unit of 1/260 mho, the characteristic admittance of the matching section ($Z_0 = 260$ ohm) used. Curve (i) of this figure shows the admittance of the slot alone; it will be seen that the impedance is capacitive above the resonant frequency and inductive below it, a behaviour opposite to that of a quarter-wave dipole. Curve (ii) shows the admittance as transformed by the quarter-wave of $Z_0 = 260$-ohm matching section, and we see that at this point the sign of the reactance variation is reversed (i.e., inductive above resonant frequency). Since it is desired to have as small standing waves as possible on the feeder over the frequency band we use a method of

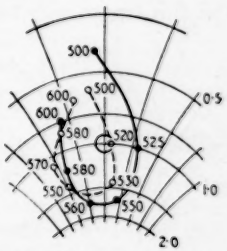


Fig. 10. *Measured admittance of slot in paraboloid; frequencies in Mc/s as indicated. Solid curve,* 13.4-*inch slot; dotted curve,* 14.6-*inch slot.*

compensating this change of impedance. A $3\lambda_0/4$ long short-circuited stub is connected across the line, its characteristic impedance being chosen so that its susceptance, which is of course zero at the mid-frequency $f_0 = c/\lambda_0$, just cancels that of the slot plus transformer at the ends of the band. This gives the result shown in the lower curve of Fig. 9, which, when transformed into 1/140 mho

units (i.e., the characteristic admittance of the main feeder), gives the upper curve. This latter curve is seen to lie entirely within the dotted circle, which corresponds to a standing-wave ratio of 1.5 : 1.

The impedance and bandwidth of the slot radiator are modified upon insertion in the paraboloid. It was found that by changing the length of the slot so as to be no longer resonant (i.e., non-reactive) at 550 Mc/s, a better overall performance could be obtained. Fig. 10 shows curves for two slots to illustrate the effect, giving

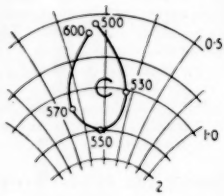


Fig. 11. *Admittance measured at compensating stub — final design; frequencies in Mc/s as indicated. This system uses a feeder Z_0 = 136 ohms; a slot 36.8-cm long and 5-cm wide; a cylinder 54-cm long and 13.4-cm in diameter; a stub 38.35-cm long having Z_0 = 49.4 ohms and a matching section 12.45-cm long having Z_0 = 262 ohms.*

the admittances at the point of attachment of the compensating stub. It will be noticed that the longer (14.6in) slot gives a smaller variation of the conductance component within the frequency range required. The longer slot also gives an inductive impedance at mid-frequency (550 Mc/s), which can be corrected by changing the length of the 260-ohm matching section or of the compensating stub or both. In fact, at this stage of the design, another parameter, the distance of the radiator from the apex of the paraboloid, was altered to improve the polar diagram slightly. This affects the impedance of the slot in a manner which may be predicted from the curve given by Westcott and Goward.[4] In the final design the slot length was 14.5in and the length of the

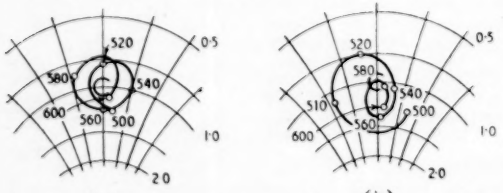


Fig. 12. *These diagrams show the measured admittance of supposedly identical slot-feeder assemblies (reproduced by courtesy of the I.E.E.). They indicate admittance* × 1/135 *mho.*

compensating stub was also altered to centre the curve on the diagram. Fig. 11 gives the final curve so obtained, together with the constants of the system.

As mentioned above, the final system was constructed with polythene cables and moulded connections at the slot. This altered the impedances a little, and the subsequent correction made use of the slight capacitive reactance introduced by one of the r.f. plugs on the feeder.[4] However, the main new feature was the fact that the constancy of characteristic impedance of the twin cable used was barely good enough for our purpose, and different slots apparently identically constructed gave different results when measured at the transmitter end of the cables, due to variation in the cable constants. A typical pair of results is given in Fig. 12 (reprinted from the paper by Westcott and Goward already cited).

5. Acknowledgments

The authors wish to acknowledge much helpful advice and discussion with colleagues at Telecommunications Research Establishment. All the diagrams of this paper are Crown Copyright Reserved, and are reproduced with the permission of the Controller of H.M. Stationery Office.

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DEPARTMENT OF SCIENTIFIC AND INDUSTRIAL RESEARCH

Sir Ben Lockspeiser has been appointed Secretary to the Committee of the Privy Council for Scientific & Industrial Research in place of Sir Edward Appleton, K.B.E., K.C.B., who relinquishes the appointment on 30th April, 1949. Sir Ben Lockspeiser is at present Chief Scientist, Ministry of Supply.

RADIO VALVE DATA

This 80-page booklet contains the characteristics of 1,600 British and American receiving-type valves. The data is tabulated and grouped under the headings: Frequency-changers, Screened Tetrodes and Pentodes, Output Valves, Diodes, Triodes, Valve and Metal Rectifiers, C.R. Tuning Indicators, Barretters and Voltage Stabilizers.

Under each main heading the British valves are grouped under manufacturers' names and are further sub-divided into obsolete, replacement and current types.

The booklet is the post-war successor to the well-known *Wireless World* Valve Data Supplements which appeared annually before the war. Compiled by *Wireless World* it is published by Iliffe & Sons, Ltd., Dorset House, Stamford St., London, S.E.1, and costs 3s. 6d. (postage 3d.).

ROTATING H-ADCOCK DIRECTION FINDER

System for High Frequencies

By B. G. Pressey, Ph.D., M.Sc.(Eng.), A.M.I.E.E.

(Communication from the National Physical Laboratory)

SUMMARY.—The paper describes a practical rotating H-Adcock direction finder of small dimensions for the frequency range 4–30 Mc/s. The tuned aerial system, which has an overall dimension of approximately 4ft (1.22 metres), is inductively coupled to the receiver through a secondary tuned circuit. Its chief features are spheres fixed at the ends of the aerials to increase their capacitance and effective height and the measures taken to balance the aerial system and to maintain that balance over the frequency range. It is shown that by 'splaying' the points of connection of the aerial coil to the horizontal feeders an appreciable reduction in polarization error may be effected.

The pick-up factor varies between 0.5 and 4 metres over the frequency range and the minimum-usable field strength varies between 8 and 2.5 μV/m. The instrumental error is not greater than 0.5° and the polarization error is of the same order as that of a U-type Adcock system. As standard-wave error is not considered applicable to elevated systems the ratio of the horizontal to the vertical pick-up factors (p.u.r.) is suggested as an alternative. The p.u.r. of this instrument varies from 0.02 to 0.13 over the frequency range.

Two models of the direction finder have been built, one a portable instrument covering the whole frequency range and the other a fixed instrument covering a restricted range (10–20 Mc/s). A feature of the latter model is the provision of an observer's seat which can be rotated with the aerial system.

1. Introduction

SOME years ago there arose a need for a high-frequency direction finder of small dimensions in connection with an investigation being conducted at the National Physical Laboratory. The instrument developed has since been used by other organizations for many purposes and as it has proved to be of general interest it is thought desirable to publish a description, though certain aspects of the design are somewhat out-of-date compared with modern techniques.

The primary considerations underlying the design of the direction finder were that it should be of small dimensions and yet have an adequate pick-up factor and reasonable immunity from polarization error. It was decided that the rotating H-Adcock type of instrument would best fulfil these requirements. Two models were built, one a fixed system covering a frequency range of 10–20 Mc/s, and the other a portable system covering the range 4–30 Mc/s. The electrical systems of the two are similar, except for the provision of additional tuning ranges on the latter.

2. General Principles of Design

Some of the principles of design of H-Adcock systems have already been set out in a paper by W. Ross and R. E. Burgess[1]. The development of the present direction finder was carried out before many of these principles were recognized and, in fact, some of the experimental work on which their paper is based was carried out on this instrument. For this reason some of the features of the present design do not conform to the principles set out by Ross and Burgess.

It has been shown[1] that for maximum sensitivity the ratio of the aerial capacitance to the total parallel aerial-circuit capacitance must be as high as possible and that the effective length of the dipoles must also be as large as possible. These requirements were met by end-loading the dipoles, thereby obtaining an increase of both aerial capacitance and effective height, and by reducing as far as possible the other capacitances of the aerial circuit. The aerial system was coupled to the receiver by means of a transformer with both primary and secondary tuning. Although it has been shown theoretically that the use of a tightly-coupled transformer with secondary tuning only is capable of giving a higher pick-up than the double-tuned type, the latter was adopted on the grounds that the loose coupling involved would greatly facilitate coil switching and screening and balancing of the primary circuit. In order to obtain high efficiency in voltage transfer from aerial transformer to receiver, intermediate coupling circuits were omitted and the secondary coil was connected directly to the grid of the first amplifier valve of the receiver, as shown in Fig. 1.

MS accepted by the Editor, November 1947

The chief mechanical considerations were portability and ease of operation. It was decided that a dimension of about 4ft* would be suitable for both the length of the aerials and the spacing between them and that the receiver should be of the light-weight battery-operated type. The second objective was achieved by reducing the inertia of the rotating portion to a minimum, providing a large easily-read bearing scale and, in the fixed model, providing a seat for the observer which could be rotated about the axis of the aerial system.

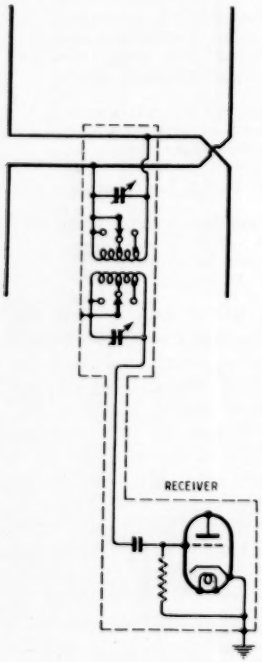
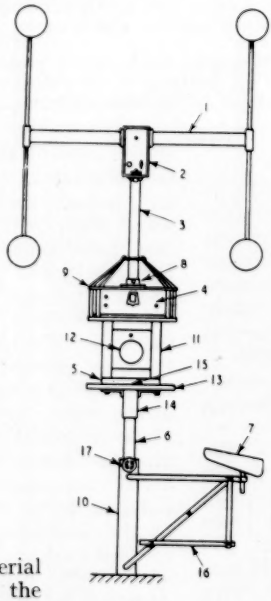


Fig. 1 (left). *Aerial circuit of the direction finder.*

Fig. 2 (below). *Fixed model of the direction finder.*



3. General Description

The circuit diagram of the aerial system is shown in Fig. 1 and the general external appearance of the fixed model and of the portable model is shown in Figs. 2 and 3, respectively. The aerial system consists of two dipoles 4ft long and spaced 4ft apart. To the ends of each are attached 6-inch diameter spheres. The dipoles are coupled in anti-phase by a twin-wire feeder, enclosed in the horizontal beam, (1, Fig. 2). Across the centre of the feeder is connected the primary coil of the aerial transformer which is contained in the aerial tuning unit (2). Both the primary and secondary coils are tuned and the signal voltage developed across the secondary is fed through a concentric feeder (3) to the input of the receiver (4). The secondary circuit, the feeder, the receiver, the battery box and the battery leads are completely screened. The aerial system, receiver and batteries are rigidly mounted on a turn-table (5) and swing as one unit when bearings are being taken. The base of the turn-table, upon which a protractor scale is mounted is supported in the fixed model by a steel column (6) around which the observer's seat (7) rotates. In the portable model the turn-table is mounted on a tripod.

4. Details of Construction

4.1 *The Aerial System*

The system was designed so that the feeder capacitance, tuning capacitance and stray capacitances across the coil were as small as possible. The feeders,

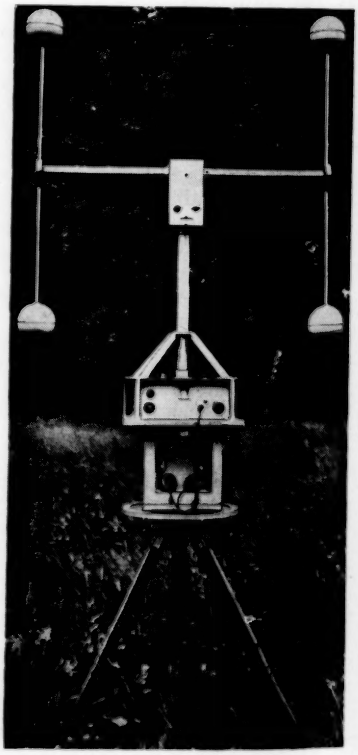


Fig. 3. *Portable model.*

* The British system of units has been used in several instances in this paper. The following are the conversion factors required to express the appropriate quantities in metric units:—

1ft = 0.3048 metre 1 lb = 0.4536 kilogram

Dimensions corresponding to standard wire gauge numbers are as follows:—16 s.w.g. = diameter 0.064 inch; 18 s.w.g. = diameter 0.048 inch; 22 s.w.g. = diameter 0.028 inch.

consisting of two 16 s.w.g. copper wires 4-ft long and spaced 1-in apart, are enclosed by a 2½-in diameter tube of non-conducting material. Both the feeder wires are in the same horizontal plane, this giving a symmetrical connection to the aerials. It is important that the cross connection of the feeder wires should be effected at their ends, as described above, rather than at the centre, since it can be demonstrated that, with the latter arrangement inequality in magnitude and phase between the e.m.fs induced in the two feeder wires is a source of polarization error. The aluminium aerial rods which are ½-in in diameter and 1-ft 9-in long are screwed into brass blocks fixed at the ends of the tube. Straps connect these blocks to the feeder terminals. The feeder tube passes right through the screening box of the aerial-tuning unit and from the mid-points of the two feeders connections are made to two studs on the outside of the tube. Actual contact to the feeder wires is made through small collars which may be moved short distances along the wires for purposes of aerial balance adjustment. Access to these collars is obtained through a window cut in the side of the tube. The ease with which rapid swings of small amplitude may be made is dependent upon a low moment of inertia of the rotating portion. The use of aluminium for the aerials in place of copper reduced this moment by one half.

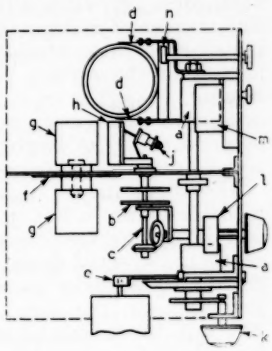


Fig 4. *Side elevation of aerial tuning unit.*

The aerial tuning unit (Fig. 4) consists of the transformer (g), tuning capacitors (a), the coil range switch (b), together with its operating mechanism (c), all of which are mounted upon the front panel of the screening box. Connection to the studs on the horizontal feeder tube is made through two German-silver strips (d) mounted on the primary tuning capacitor, and connection to the vertical feeder by means of a clip (e) mounted on the secondary capacitor. Thus the complete unit can be readily withdrawn from the screening box for inspection or adjustment.

The primary and secondary circuits of the aerial transformer are completely screened from each other and between the two coils is a radial-type electrostatic screen (f). In the tuning unit for the fixed model the frequency band of 10–20 Mc/s is covered in three coil ranges and the 4–30 Mc/s band of the portable model is covered in six ranges. It would have been possible to cover the band in fewer ranges but this would have necessitated the use of a considerably larger tuning capacitor and, in consequence, a smaller pick-up factor would have been obtained at some frequencies. The coils (g) are wound on 1½-in diameter Calit formers with 18 s.w.g. copper wire. The smaller values of inductance required for the higher-frequency ranges are obtained by shorting turns of the coils at the ends furthest from the electrostatic screen so that the coupling is maintained at approximately its critical value.

The range switches (h and b) are ganged and controlled through gearing by a knob on the front panel. The switch (b) on the secondary side is of the low-capacitance multi-position type. The switch (h) on the primary side is of special design. As will be shown later it is essential that there should be negligible capacitance (<0.1 pF) between the contacts and any earthy point. This then, was the primary consideration affecting the switch design. The low contact capacitance is obtained by attaching the contacts directly to the coil and mounting the wipers on a section of ebonite cylinder so that as it is rotated the contacts are shorted by the wipers. Constructional details of the three-range switch are shown in Fig. 5. The construction of the six-range switch is similar. Preset balancing capacitors of 12-pF maximum capacitance (j) are connected between each of these wipers and the earthed operating spindle.

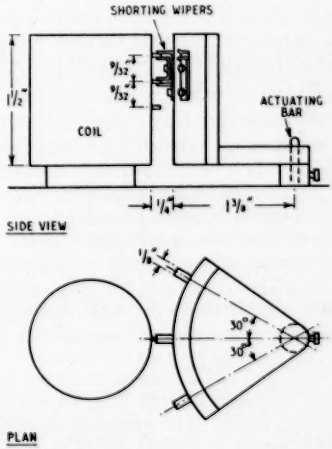


Fig. 5. *Coil range switch.*

The two ganged tuning capacitors (a, Fig. 4) which have a maximum capacitance of 40 pF each, are controlled through a slow-motion drive by a knob (k) and a trimming capacitor (l) is provided on the secondary side. An aluminium screen (m) is placed round the moving plates of the primary capacitor and connected to them in order that their capacitance to the screening box might not change with their rotation. The importance of this precaution is explained later.

A device for the determination of sense is fitted on the fixed model. The principle upon which it operates is that of temporarily making the aerial asymmetrical and so producing a reciprocal difference. The direction of the resultant change in bearing enables the observer to determine the sense. Aerial asymmetry is obtained by introducing a non-inductive resistor of 8.5 ohms between the upper feeder contact strip (d) and the lead to the coil (g). Across the resistor is a low-capacitance shorting switch (n) operated from the front of the panel.

The concentric vertical feeder (3, Fig. 2) connecting the secondary circuit to the input of the receiver is designed so as to have a low capacitance. It consists of a single 22 s.w.g. copper wire running down the axis of a 2-in diameter brass tube. The capacitance of this 2-ft long feeder is 12 pF. At the lower end of the feeder wire is a plug fitting into the input socket of the receiver. A brass cup (8) which receives the base of the screening tube ensures good screen contact. A slot in the side of this cup and a key attached to the tube prevent rotation of the tube relative to the receiver.

4.2 *The Receiver*

The receiver is of a standard commercial pattern and is operated from batteries. The control grid of the r.f. amplifying valve is connected through a coupling capacitor to the input socket and modifications have been made in order to reduce the input capacitance to a minimum. The reduction of this capacitance is important since it affects the tuning range of the secondary circuit of the aerial transformer.

4.3 *Supporting Structure*

The receiver and aerial system are supported by a wooden framework (11, Fig. 2) mounted on a turn-table. A screened battery box, a loud-speaker (12) and phone jack, and a container for the receiver tuning coils are fitted inside this framework. The base of the turn-table carries a 15½-in diameter protractor scale (13) with which are associated two pointers (15) fixed to the front and back of the rotating top of the turn-table.

In the fixed model the turn-table is mounted on a 2½-in diameter steel column, the base of which is firmly imbedded in the ground. The observer's seat is attached to a tube (10) which is capable of rotation about this column. A platform (16) is provided so that the observer when standing upon it may conveniently reach the aerial-tuning controls. A turn of a small hand-wheel (17) clamps the seat in any position. The purpose of making the seat rotatable is that the observer may quickly and accurately place himself in the correct position when taking a bearing. As will be explained later, the observer's position is of great importance for accurate observations. The height of the centre of the aerial system is 8½ft from the ground.

The wooden tripod which supports the portable model is shown in Fig. 3, and a 'shooting-stick' seat is used for the observer. The height of the centre of the aerial system is 7½ft. For transport the direction finder is readily dismantled into four sections, the aerial rods, the transformer box with the horizontal and vertical feeders, the receiver and battery box, and the tripod. The total weight of the equipment with batteries is 75 lb.

5. Development

Antenna effect, the presence of which was shown by flatness of bearing and reciprocal difference, was eliminated by means of a trimming capacitor connected between one end of the aerial coil and the screening box. The cause of this antenna effect and the reason for its elimination by this method is given in Appendix I. The adjustment of this trimmer was very critical, balance to within about 0.1 pF being required for the best performance of the system.

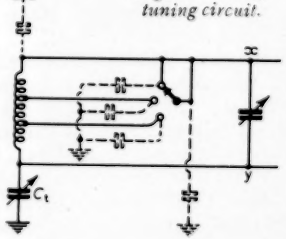


Fig. 6. *Aerial tuning circuit.*

It was found that the adjustment would only hold at any one wavelength. Thus if the system was balanced on the mid-frequency of one of the ranges so that the reciprocal difference was less than 0.5°, the difference at either end of this same range was as large as 6°. Both the coil range switch and the tuning capacitor were found to be responsible.

The original range switch used on the aerial coil was of the low-capacitance multi-position type. The circuit diagram of the coil and switch showing the switch capacitance is given in Fig. 6. The effect of the adjustment of the trimmer C_t is to equalize the currents flowing from points x and y to earth including those through the switch-contact capacitances. These contact capacitances are small in value (1 pF each) but since they are in series with sections of the coil any balance obtained in this way can hold at only one frequency. If the switch is to be satisfactory the capacitance of its contacts must be negligible (<0.1 pF). This was achieved by the new type of switch described in the previous section, in which the contacts are fixed directly on the turns of the coil.

Originally the tuning capacitor used was of the 'series gap' type with the rotating plates earthed. If the capacitance to earth of the two ends of the aerial coil must be balanced to within 0.1 pF, it follows that the capacitances of the two halves of this capacitor must also be balanced to the same degree for all positions of the rotating plates. The best balance that it was possible to obtain by adjustment of the plates was within 0.5 pF only. The 'series gap' capacitor was therefore abandoned in favour of a single capacitor with the moving plates enclosed by a fixed screen connected to them.

These changes to the coil switch and the tuning capacitor resulted in a great improvement over the whole band in the sharpness of bearing (average swing on local transmitter $= \pm 0.5°$) and in a reduction of reciprocal difference to less than 0.5°.

Preliminary measurements of polarization error on the fixed model at a frequency of 15 Mc/s indicated that it was of the order of $2\frac{1}{2}°$ for an angle of incidence of 64° and an angle of polarization of 45°. At the time of these measurements, connection from the contact studs on the feeder tube was made to the centre points of the feeders. As shown in Appendix I, compensation for the asymmetry of the aerial system, to which the error is due, may be effected by alteration of the position of these tapping points. Movements of these points in opposite directions along the feeder wires for a distance of 1½in resulted in a decrease of error to 1½°. It was found, however, that this improvement was not obtained at other frequencies and that readjustment of the taps was necessary for optimum performance. This was particularly so on the portable model, in which a greater frequency range was covered, and at the highest frequency (30 Mc/s) the adjustment was very critical, a movement of ½in resulting in a change of error of 2° or 3°. However, when the optimum position for the whole frequency range had been found the overall performance showed a distinct improvement on that obtained with central tapping points.

The polarization error was influenced to a certain extent by the position of the observer; i.e., by whether he was facing or had his back to the transmitter. The magnitude of this effect was proportional to the degree of asymmetry of the aerial system and of flatness of the bearing. However, its influence on the polarization error measurement could be eliminated by taking the mean of the errors measured in the two positions of the observer.

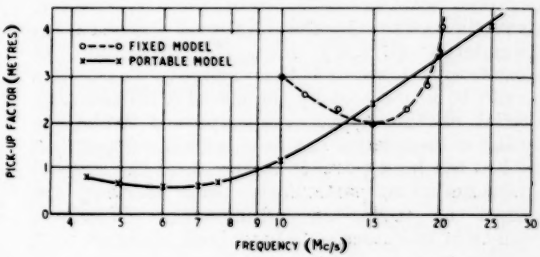


Fig. 7. *Pick-up factor of aerial system.*

6. Performance

6.1 *Sensitivity*

The pick-up factor of the aerial system was obtained from the measured value of the aerial output voltage (i.e., the voltage across the lower end of the vertical feeder) and the calculated value of the incident field producing this output. The output voltage was determined by a substitution method using a standard-signal generator as the calibrated voltage source and the receiver as the indicator. The field was produced by a local transmitter with a loop aerial containing a current meter.

The results of these measurements on both models are shown in Fig. 7. The increase in pick-up at the high-frequency end of the bands is due to the increase of aerial spacing factor and of effective height and the increase at the low-frequency end is probably due to an increase in the magnification of the tuned circuits.

The measured increase in pick up due to the aerial spheres was approximately three times. This is in agreement with the theoretical increase in the aerial capacitance of 2 to 1 and in the effective height of 1.5 to 1.

Although the pick-up factor measures the efficiency of the aerial system from the voltage transfer point of view, it does not indicate the

ultimate sensitivity of the direction finder in operation as the presence of circuit noise has been neglected. As a more practical indication of sensitivity the minimum field strength required to give a bearing of reasonable accuracy was measured. In order to obtain a bearing on the fixed model with a swing of $\pm 5°$ the field strength required was 2.5 μV/m at a frequency of 15 Mc/s and the corresponding figure for the portable model was 2 μV/m at 25 Mc/s and 4 μV/m at 10 Mc/s.

6.2 *Polarization error*

For the first measurements of polarization error and investigations into its reduction by adjustment of feeder taps, a test transmitter with a dipole aerial was used. Subsequently, however, investigations at the National Bureau of Standards (U.S.A.) into the technique of measurement showed that serious inaccuracies could be introduced by the use of a transmitting aerial of this type. An improved technique using a loop aerial has now been developed but it has not been possible to repeat all the original measurements, particularly those showing the effect of feeder-tap adjustment. However, sufficient measurements have been made on both models to show the order of the performance.

The test transmitter was situated at the top of a 72-ft tower at a distance of 190ft from the direction finder. The angle of polarization of the wave was 84°. The method of procedure was the same as that already described in a previous paper[2]. Both the bearing error and the swing were measured so that from them could be calculated the total error, which is the error obtained when the wanted and unwanted pick-ups are in phase. The measured errors were converted to equivalent errors for 45° polarization and are plotted in Fig. 8.

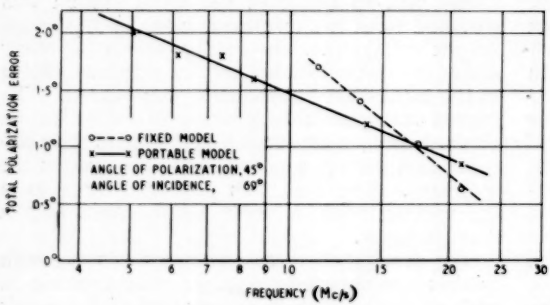


Fig. 8. *Polarization error.*

It was decided to present the results in the above way rather than to convert them to standard-wave errors as such a figure of merit is not altogether satisfactory for an elevated system of this type. The s.w.e. is dependent upon the height of the system from the ground, particularly if this exceeds a quarter of a wavelength, and the polarization error for an angle of incidence other than 45° cannot be readily computed from it. Another figure of merit is suggested; viz., the ratio of the horizontal to vertical pick-up factors (p.u.r.). This has the advantage of being more fundamental to the system and is substantially independent of height, except when the aerials are very near the ground. The p.u.r. values have been calculated for a number of frequencies by the method described in Appendix II and are given in Table I.

TABLE I

Frequency (Mc/s)	P.U.R. of Fixed Model	P.U.R. of Portable Model
6		0.13
8		0.09
12	0.06	0.05
20	0.02	0.02

The values of polarization error given are probably pessimistic ones since, as mentioned above, there was no opportunity of checking that the aerial system was in the optimum condition before the measurements were taken. An improvement of as much as 2 to 1 might be obtained by adjustment of the feeder taps. Nevertheless, the figures given are of the same order as those for a buried U-Adcock system on a good site. It has been found[1] that a large portion of the polarization error is due to the presence of the receiver between the lower aerials partly on account of the increased capacitance between them and partly on account of its re-radiated field.

6.3 *Observations on distant stations*

A short series of observations was made with the fixed model on distant stations in order to test the instrument as a practical direction finder. A total of 75 bearings was obtained on 40 different stations situated at distances up to 7,000 miles. The bearing obtained in each case was the mean of five separate observations. All the mean bearings were within 7.5° of the true bearing while 31% were within 1°, 45% within 2°, 65% within 3° and 91% within 5°. The results of these observations compare very favourably with similar observations taken on other types of Adcock system.

6.4 *Operation*

The operation of the direction finder is generally

satisfactory. The turn-table rotates very freely and, in spite of the weight of the system, rapid swings of small amplitude may be made without difficulty. The adjustments to the aerial-balancing capacitor and to the feeder taps are initial ones and require little further attention after the instrument has been put into service. Of the two the balancing capacitor is the more likely to drift out of adjustment but this can readily be checked by the measurement of the reciprocal difference. This difference does not normally exceed 0.5° on the fixed model and 1° on the portable model.

Operation of the sense switch on the fixed model produces a change in bearing of 10–13 degrees according to frequency. There is also a small decrease in sharpness of the bearing, particularly on the higher frequencies. Sense determination on signals from distant stations is quite definite provided that the swing is not greatly in excess of $\pm$ 10°.

The position of the operator has a marked effect upon the sharpness of the bearing. It is necessary for him to be placed in exactly the minimum position and for his head to be kept central. Although a slight inclination of his head may increase the flatness of the bearing, an appreciable movement is required in order to produce a noticeable error.

7. Conclusions

The performance of the direction finder has shown that it is possible to develop an instrument of small dimensions which has a comparatively high sensitivity. By means of suitable adjustments on the aerial system the instrumental errors can be reduced to a low value so that its accuracy is comparable with that of the commonly used U-Adcock system. It has also been shown how critically dependent the instrumental error is on the balance of the aerial system in relation to both the aerial circuits and the position of the observer.

The performance of the direction finder may be summarised as follows :—

(i) The sensitivity is such that on the fixed model over its restricted range of 10–20 Mc/s bearings with a swing of $\pm$ 5° may be taken on stations giving a field strength as low as 2.5 μV/m ; and on the portable model the corresponding minimum field strength varies between approximately 8 μV/m and 2 μV/m over the frequency range 4.5 to 25 Mc/s.

(ii) The instrumental error when receiving vertically-polarized radiation is not greater than 0.5°.

(iii) The pick-up ratio varies between 0.02 at 20 Mc/s and 0.13 at 6 Mc/s which results ie polarization errors of the same order as thos obtained with buried U-Adcock systems used under the same conditions.

8. Acknowledgment

The work described in the paper was carried out as part of the programme of the Radio Research Board, to whom it was communicated in a confidential report dated February 1939. It is published by permission of the Department of Scientific and Industrial Research.

The author desires to acknowledge the valuable guidance given by Dr. R. H. Barfield during the course of the work.

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[2] R. L. Smith-Rose and W. Ross: "The Use of Earth Mats to Reduce the Polarisation Error of U-Type Adcock Direction Finders," *J. Instn elect. Engrs*, 1947, Vol. 94, Pt. III, p. 91.

APPENDIX 1

Errors due to Electrical Asymmetry of the Aerial System

The condition for the system to be free from error is that there shall be no current in the aerial coil when the e.m.fs induced in the aerials are co-phasal ; i.e., when the plane of the aerials is perpendicular to that of propagation. The errors may be divided into two sections, those due to antenna effect and those due to horizontal pick-up of the system. The cause and method of elimination of some of these errors may be determined by a detailed consideration of the system.

In Fig. 9(a) is shown the circuit of the aerial system including the capacitances to earth of the aerials and of the tuning circuit. The instantaneous directions of the e.m.fs induced in the system by vertical and horizontal fields are indicated by arrows. Fig. 9(b) gives the equivalent circuit of the system, the feeders being represented by series impedances $Z_1 \ldots Z_4$ and the e.m.fs by equivalent generators. On account of the symmetrical construction it may be assumed that the magnitudes of the e.m.fs induced in the aerials are equal to each other and that the four sections of the feeders have the same impedance. The capacitances of the aerials are not all equal owing to the fact that the lower pair are nearer the ground than the upper, but it may be assumed that the two top ones are equal to each other and that the two bottom ones are also equal.

Consider first the errors due to antenna effect, for which purpose $\mathfrak{E}_h$ may be assumed to be zero and hence

$$e_1 = e_2 = e_3 = e_4 = 0.$$

If the current flowing round circuit 1 (Fig. 9) is equal in magnitude and in phase to that round circuit 4 then there is no resultant current in C_a. Similarly zero current in C_b is produced by equality between the currents in circuits 2 and 3. However, as $C_1 \neq C_4$ and $C_2 \neq C_3$, there is a resultant current in C_a and C_b. But, if $C_1 = C_3$ and $C_2 = C_4$ and C_a is made equal to C_b the potentials of points 'a' and 'b' with respect to earth will be equal, and there will be no current in the aerial coil. It is to be noted that if all the aerials have equal capacitances there will be no antenna effect and hence no need for adjustment of C_a and C_b.

Now consider the errors due to horizontal pick-up. Owing to the inequalities in aerial capacitances there are additional resultant currents flowing through C_a and C_b

due to the e.m.fs e_1, e_2, e_3 and e_4 induced by the horizontal field $\mathfrak{E}_h$. As points 'a' and 'b' are at the centre of the feeders $e_1 = e_4$ and $e_2 = e_3$. These additional currents, however, flow between 'a' and 'b' in the same direction and thus give rise to a potential difference between these points and, hence, to current in the aerial coil. Therefore it is necessary that these resultant currents should be

ψ = phase angle change on reflection
ϕ = phase angle due to path difference between direct and reflected waves.

In both cases the surface-wave component has been neglected since, for the conditions of measurement, its magnitude is less than one-tenth of the space-wave component.

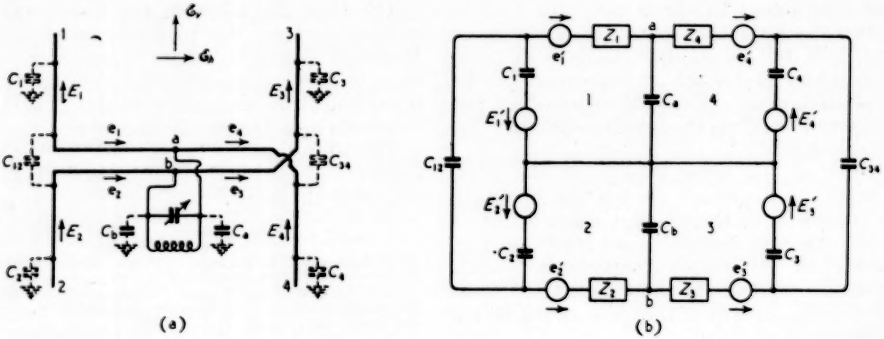


Fig. 9. *Actual and equivalent aerial system.*

reduced to zero. This can be done by changing the magnitudes of the e.m.fs e_1, e_2, e_3 and e_4 and of the impedances Z_1, Z_2, Z_3 and Z_4 by moving the tapping points 'a' and 'b' along the feeders. If 'a' is moved so that e_1 and Z_1 are decreased and e_4 and Z_4 increased, and 'b' moved so that e_3 and Z_3 are decreased and e_2 and Z_2 increased, the resultant currents through C_a and C_b may be reduced to zero.

In the case considered above the feeders are unscreened so that the e.m.fs due to the horizontal field are induced directly into the feeder wires. If the feeders are screened the current induced in the screening tube will give rise to a re-radiated field which will induce e.m.fs in the vertical aerials. The equivalent circuit diagram will still apply, however, and the same conclusions will be obtained, except that movement of the tapping points will not change the magnitudes of the e.m.fs.

APPENDIX II

Calculation of Horizontal to Vertical Pick-up Ratio

Let the vertical and horizontal components of the field from the transmitter be $\mathfrak{E}_v$ and $\mathfrak{E}_h$ respectively and those of the resultant field at the aerial system be E_v and E_h respectively (see Fig. 10).

Then $E_v = \mathfrak{E}_v \sin^2 i[1 + \rho_v \angle \phi + \psi_v]$

and $E_h = \mathfrak{E}_h[1 + \rho_h \angle \phi + \psi_h]$

where ρ = reflection coefficient

i = angle of incidence of the wave

Horizontal pick-up $= E_h \, x \cos \epsilon$

Vertical pick-up $= E_v \, y \sin \epsilon \sin i$

where ϵ = angle of error

x and y = horizontal and vertical pick-up factors of the system.

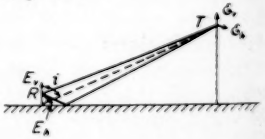


Fig. 10. *Position of transmitter and receiver for polarization measurements.*

The sharpness of the bearing depends upon the relative phases of the wave components and of the pick-up factors. If the phases are such that the minimum is silent, then ϵ is equal to the total polarization error and we may equate the horizontal and vertical pick-ups to give the pick-up ratio as,

$$\text{P.U.R.} = \left|\frac{x}{y}\right| = \left|\frac{E_v}{E_h}\right| \tan \epsilon \sin i$$

Assuming ground constants of

$$k = 20 \text{ and } \sigma = 3 \times 10^8 \text{ e.s.u.}$$

the values of E_v and E_h were calculated and from the measured values of polarization error the values of the p.u.r. were determined.

RESISTIVE FILMS IN VALVES

Effect on Interelectrode Capacitance

By E. G. James, B.Sc., Ph.D., A.M.I.E.E. and B. L. Humphreys, B.Sc.

(Communication from the Research Staff of the M.O. Valve Co., at the G.E.C. Research Laboratories, Wembley, England)

SUMMARY.—This paper applies Professor Howe's analysis of the behaviour of resistances at high frequencies, to the behaviour of interelectrode capacitances in the presence of resistive films between the electrodes. It is shown that the capacitance of such films varies inversely as the square root of the frequency at high frequencies and tends to a limit at low frequencies. The importance of the frequency in the measurement of interelectrode capacitances is demonstrated experimentally.

1. Introduction

In some measurements of the interelectrode capacitances of radio valves, it has been noticed that the capacitance measured with an audio-frequency capacitance bridge, is frequently higher than that measured with a capacitance bridge operating at radio frequency. This dependence of the measured capacitance upon the frequency of measurement, has been found to be associated with the presence of resistive films on the insulators supporting the valve electrodes. Such films may be caused by deposition of getter or by the evaporation of metal from the cathode either during the manufacture or during the life of the valve. The value of the resistance may be as high as 1000 megohms or more, and is often sufficiently high not to affect the performance of the valve at radio frequency. The measured value of the capacitance is, however, modified by the presence of the resistance and cannot be used to predict the performance of the valve at frequencies higher than that of the measurement.

In his Editorial in *Wireless Engineer* of June 1935, Professor Howe presents an analysis explaining the decrease of the effective resistance of a resistor with increase of frequency. This analysis involves the assumption that the resistor behaves as a transmission line whose characteristics are determined by the resistance and capacitance per unit length of the resistor. The method of calculating the capacitance per unit length of a resistor, having a diameter small compared with its length, is given in this Editorial and the method for a resistor whose diameter is comparable with its length is given in the Editorial of November 1940.

2. Analysis

Following Professor Howe's arguments, a resistor of length $2l$ may be considered as a transmission line of length l whose impedance Z, from the sending end, is given by the equation

$$Z = Z_0 \tanh (\alpha l) \quad .. \quad .. \quad .. \quad (1)$$

In this equation

$$Z_0 = \sqrt{\frac{R_0}{j\omega C_0}}$$

and $$\alpha = \sqrt{j\omega C_0 R_0}$$

where R_0 and C_0 are respectively the resistance and capacitance per unit length of the equivalent transmission line.

Inverting both sides of equation (1), we have

$$Y = \frac{1}{Z_0} \coth (\alpha l)$$

$$= \frac{\alpha}{R_0}\left(\frac{e^{\alpha l} + e^{-\alpha l}}{e^{\alpha l} - e^{-\alpha l}}\right) \quad .. \quad .. \quad (2)$$

2.1 *Small Values of* αl

Expanding and simplifying the R.H.S. of equation (2) gives

$$Y = \frac{\alpha}{R_0}\left[\frac{1 + (\alpha l)^2/2! + (\alpha l)^4/4! + (\alpha l)^6/6! + \ldots}{\alpha l\{1 + (\alpha l)^2/3! + (\alpha l)^4/5! + (\alpha l)^6/7! + \ldots\}}\right].$$

Writing $(\alpha l)^2 = jA$, where $A = \omega C_0 R_0 l^2$,

$$Y = \frac{1}{R_0 l}\left[\frac{1 - A^2/4! + \ldots + j\{A/2! - A^3/6! + \ldots\}}{1 - A^2/5! + \ldots + j\{A/3! - A^3/7! + \ldots\}}\right]$$

$$= \frac{1}{R_0 l}\left[\frac{1 + A^2/30 + \ldots + jA\{\frac{1}{3} + A^2/630 + \ldots\}}{1 + A^2/90 + \ldots}\right]$$

$$\quad .. \quad .. \quad .. \quad .. \quad (3)$$

When ω is so small that second and higher order terms in A are negligible, equation (3) reduces to

$$Y = \frac{1}{R_0 l}\left[1 + j\frac{A}{3}\right]$$

and putting $Y = G + j\omega C$

$$G = 1/R_0 l$$

and $$C = C_0 l/3$$

which, of course, are in agreement with the calculations made in the Editorial, the resistance tending to the d.c. value of the resistor, and the parallel capacitance to one-third of the product of the length and the capacitance per unit length of the equivalent transmission line. The latter

MS accepted by the Editor, January 1948

was pointed out by Hartshorn in his paper to *Wireless Engineer*, July 1938.

The approximation for the conductance is correct to within 5% provided that

$$A^2/30 < 0.05$$

i.e., $\omega C_0 R_0 l^2 < \sqrt{1.5}$

or $\omega C/G < 0.41$

and that for the capacitance, provided

$$A^2/630 < 0.05/3$$

i.e., $\omega C_0 R_0 l^2 < \sqrt{10.5}$

or $\omega C/G < 1.1$

2.2 *Large Values of* αl

The equation (2) may also be written

$$Y = \frac{\alpha}{R_0}\left(1 + \frac{2}{e^{2\alpha l} - 1}\right) \quad .. \quad .. \quad (4)$$

Now $$\alpha = \sqrt{j\omega C_0 R_0}$$

$$= \sqrt{\frac{\omega C_0 R_0}{2}}\left(1 + j\right)$$

and if $\frac{1}{e^{2\alpha l} - 1} = a + jb$,

equation (4) becomes

$$Y = \sqrt{\frac{\omega C_0}{2R_0}}\left(1+j\right)\left\{1+2(a+jb)\right\}$$

$$= \sqrt{\frac{\omega C_0}{2R_0}}\left[1+2(a-b)+j\{1+2(a+b)\}\right]$$

$$.. \quad .. \quad (5)$$

When ω is sufficiently large, both a and b will be small compared with unity and (5) reduces to

$$Y = \sqrt{\frac{\omega C_0}{2R_0}}\left(1 + j\right)$$

and putting $Y = G' + j\omega C'$

$$G' = \sqrt{\frac{\omega C_0}{2R_0}}$$

and $$C' = \sqrt{\frac{C_0}{2\omega R_0}}.$$

These approximations will be correct to within 5% provided both $|a - b|$ and $|a + b|$ are less than 0.025.

Now $$a + jb = \frac{1}{e^B(\cos B + j \sin B) - 1}$$

where $$B = l\sqrt{2\omega C_0 R_0}$$

and on rationalizing,

$$a + jb = \frac{e^B \cos B - 1 - je^B \sin B}{e^{2B} - 2e^B \cos B + 1}.$$

Therefore

$$a - b = \frac{e^B(\cos B + \sin B) - 1}{e^{2B} - \ldots + 1}$$

and $$a + b = \frac{e^B(\cos B - \sin B) - 1}{e^{2B} - 2e^B \cos B + 1}$$

Thus both $|a + b|$ and $|a - b|$ are less than

$$\frac{\sqrt{2}\, e^B - 1}{(e^B - 1)^2}.$$

Hence the condition that the approximations for the conductance and capacitance should be valid to within 5% is that

$$\frac{\sqrt{2}\, e^B - 1}{(e^B - 1)^2} < 0.025$$

i.e., $$e^{2B} - 2e^B(1 + 20\sqrt{2}) + 41 > 0$$

This is true if $e^B > 57$

or $$l\sqrt{2\omega C_0 R_0} > \log_e 57$$

i.e., $$\omega C_0 R_0 l^2 > 8.25$$

or $$\omega C/G > 2.8$$

2.3. The analysis shows that to a first approximation the effective resistance and capacitance of a resistor or a resistive film, remain constant at low frequencies and are inversely proportional to the square root or the frequency at high frequencies. It will be noted that the range of values of $\omega C/G$ which is not covered by the two sets of approximations is small, so that the change of law for a particular resistor or resistive film, takes place over a comparatively small range of frequencies. The capacitance law changes in just over one octave, while that for the resistance changes in just under three octaves.

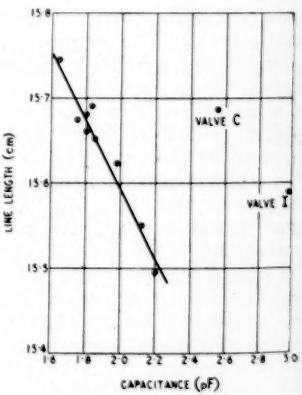


Fig. 1. *The resonant line length at 600 Mc/s v. valve capacitance.*

3. Experimental

The interest in the problem first arose from measurements of the interelectrode capacitances of certain experimental valves. These valves were such that the anode and the grid were in the form of discs and were separated by a glass

tube 19-mm internal diameter and 7-mm long. When the anode-to-grid capacitance was measured by means of a bridge operating at 1,600 c/s, certain valves appeared to have a capacitance about twice that of normal valves. This variation of capacitance was not substantiated by measurements of the line lengths required for resonance at a frequency of 600 Mc/s, when the valve was operated in a coaxial-line circuit. The valves were then measured by means of a bridge operating at 1 Mc/s and the two sets of measurements are given in Table I. The conductance measured at 1 Mc/s is also given.

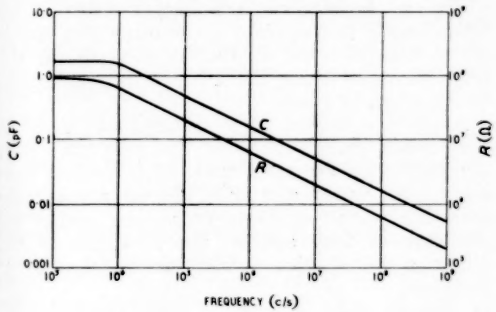


Fig. 2. *The variation of effective capacitance and resistance with frequency.*

It will be seen that when no conductance is present, the readings at the two frequencies are approximately the same, but when some conductance is present, the higher-frequency measurements are lower than the lower-frequency measurements.

Fig. 1 shows a plot of the resonant-line length at 600 Mc/s against the capacitance measured at 1 Mc/s. It will be seen that the valves having a low conductance fall on a curve, while those having a high conductance have a much higher capacitance than the line length would indicate.

Making the assumption that the resistive film was uniform between the anode and the grid of the valve, it was estimated that the capacitance per unit length of the equivalent transmission line was 14 pF per cm. The variation of capacitance and resistance of a film of this nature having a low-frequency resistance of 100 meghoms, is shown in Fig. 2. At low frequencies the calculated effective capacitance of the film is 1.63 pF, while that at 1 Mc/s is 0.08 pF and that at 600 Mc/s is 0.0033 pF. Thus the film plays little part in the effective capacitance of the valve at very high frequencies, but an appreciable one at low frequencies.

A similar effect has been noticed in the measurement of the anode-to-grid capacitances of certain pentode valves. Table II shows the comparison of measurements made at 1 Mc/s and at 1,600 c/s. It will be seen that when a resistive film is present between the electrodes, the capacitance measured at 1600 c/s is many times that measured at 1 Mc/s but when no resistive film is present, the two capacitances are approximately the same.

TABLE I

Valve	1-Mc/s Bridge		1,600-c/s Bridge	Line length for resonance at 600 Mc/s.
	Capacitance (pF)	Conductance (μ mho)	Capacitance (pF)	(cm)
A	1.64	0.15	3.15	15.74
B	1.83	0.35	3.52	15.69
C	2.75	2.85	3.97	15.69
D	1.80	0.05	1.90	15.68
E	1.75	0.0	1.85	15.67
F	1.80	0.0	1.90	15.66
G	1.85	0.0	1.90	15.65
H	1.98	0.3	2.93	15.62
I	3.18	1.7	5.00	15.59
J	2.12	0.4	4.00	15.55
K	2.20	0.0	2.20	15.49

TABLE II

Valve	1-Mc/s Bridge		1,600-c/s Bridge
	Capacitance (pF)	Conductance (μ mho)	Capacitance (pF)
1	0.0067	0.0	0.0065
2	0.0068	0.002	0.027
3	0.0060	0.006	0.036

4. Conclusions

The analysis and the experimental evidence given show the importance of the frequency of measurement when measuring capacitances which may be accompanied by resistive films. Such conditions are often encountered in valves and this paper has been written to emphasize the existence of the effect, which does not seem to have had the same universal appreciation as the associated resistance effect.

5. Acknowledgment

The Authors desire to tender their acknowledgments to The M.O. Valve Co., Ltd., on whose behalf the work described in this publication was carried out.

SPIRAL-PHASE FIELDS

Possible Applications

By E. K. Sandeman, Ph.D., B.Sc., A.C.G.I., M.I.E.E.

1. Carrier Fields

IF four vertical aerials are arranged one at each corner of a horizontal square of side equal to less than a quarter of a wavelength, and if each diagonal pair is driven in mutual anti-phase and in quadrature with the other diagonal pair, a carrier-frequency spiral-phase field is originated. Such a field was first used by Chireix to obtain low-angle radiation, free from high-angle lobes, with a structure not large compared to the wavelength; for this purpose the spiral-phase pattern produced had no significance other than that it was the concomitant of a circular polar diagram in the horizontal plane.

It is useful here to define the carrier frequency referred phase difference as the difference in phase between the carrier-frequency field at any point in space and any arbitrary standard of fixed phase at the transmitter location.

With the arrangement described above, contours of constant referred phase difference lying in the horizontal plane are substantially in the form of Archimedes spirals of pitch equal to the wavelength. An Archimedes spiral is one in which the length of the generating radius vector increases in direct ratio with the angle through which the vector turns. True Archimedes spirals are only produced when the spacing between aerials becomes zero.

Fields having the above characteristics may be called horizontal Archimedes fields.

Figs. 1 and 2 each show two sets of contours of which each set defines the lines of constant phase in the plan view of a horizontal Archimedes field. The direction of rotation of the spirals depends on the phasing of the four aerials: reversing the phase of either pair of diagonally opposed aerials reverses the direction of rotation of the field.

In Fig. 1 each spiral rotating in a clockwise direction defines a line of constant phase, constant phase spirals being drawn at intervals of 30° in phase, and the whole set of spirals defining an Archimedes field. The spirals rotating in an anticlockwise direction define an Archimedes field with the opposite sense of rotation.

It will be seen that in each field there is only a single spiral defining the line of zero phase; these fields are therefore called single-phase fields. As will be explained later, in the case of envelope fields, two-phase, three-phase, etc., fields are possible.

In a single-phase field, during one circuit of the centre a phase change of 2π occurs, in a two-phase field the phase change per circuit is 4π and so on.

Fig. 2 shows two Archimedes fields which differ widely in frequency. The pitch of a spiral-phase field is equal to the wavelength (at the frequency generating the field) in the medium in which the field exists.

A physical model of a spiral-phase field has been made using water as the medium. Four plungers were arranged with their conical ends dipping in water and were driven sinusoidally with the phase relations prescribed earlier for the case of four aerials, the locations of the plungers being also analogous to the aerial positions.

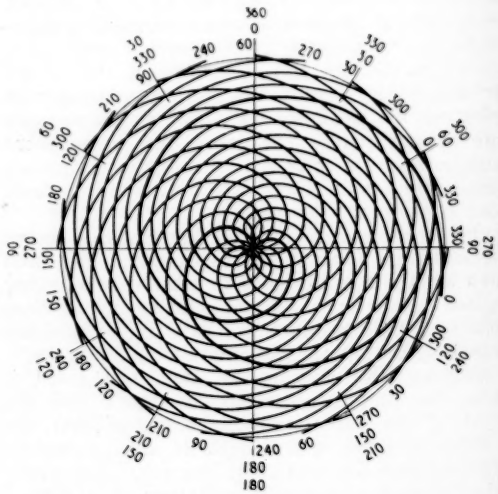


Fig. 1. *Archimedes spirals defining constant phase contours of two oppositely-directed spiral-phase fields of different pitch.*

The result was a water wave consisting of a spiral rather like the slow spiral in Fig. 2, the whole spiral rotating at the generating frequency.

If a vertical aerial, radiating the same carrier frequency as that radiated from the four aerials radiating the spiral-phase field, is set up at the

centre of the square of aerials it gives rise to a carrier-frequency field of which the contours of constant referred phase are circles, and, further, the distance between circles defining phases which differ by 2π radians is equal to the wavelength at the carrier frequency. Such a field will be called a circular-phase field. The circular-phase field set up in this way will be said to be concentric with the carrier-phase field.

The difference in phase between any two fields at any point in space will be called the local-phase difference.

The contours of constant local-phase difference between a spiral-phase field and a concentric circular-phase field are in the form of radii originating at the common centre of the fields. Such contours are obviously real and directly observable and may be used by an aircraft to home on to the common centre. The local-phase difference defined by any one of such radial contours will be called the radius phase.

With the arrangement described, the radius phase is approximately proportional to the bearing along which the radial constant-phase contour lies; evidently by adjustment of the phase reference, the radius phase can be made substantially equal to bearing. The change of radius phase observed by a craft making a circuit of the common centre is 2π radians, and a spiral-phase field in which this occurs has been called a single-phase field. Under the heading Envelope Fields, means of producing multiple fields, in which the change in phase per rotation is a multiple of 2π, will be described.

If the radius phase on bearing 0° is arranged to be 0°, then the difference between the radius phase and the bearing has a maximum deviation which is reduced as the separation between the aerials originating the spiral-phase field is reduced. If, for instance, the diagonal separation between aerials is 0.07 wavelength, the maximum deviation is less than half a degree. This deviation is constant, known and capable of correction.

2. Envelope Fields

While the concept of carrier-frequency spiral-phase fields has been used to introduce the subject, it is considered that envelope fields offer much greater promise of future application. The terms referred-phase difference, local-phase difference, and radius phase will be applied to the phase of the envelopes of fields, originated by the radiation of sinusoidally-modulated carrier frequencies, with definitions analogous to those used for carrier-frequency fields.

Spiral-phase envelope fields in which the lines of constant referred envelope phase difference are Archimedes spirals will be termed envelope Archimedes fields. In such fields, when horizontal, the envelope radius phase is proportional to the azimuth angle.

If the graph of $r = a \cos \theta$ is plotted on polar co-ordinates, the resulting curve consists of two circles of diameter a with centres on the horizontal axis, kissing one another at the origin. In general, if the graph of $r = a \cos n\theta$ is plotted, a diagram with $2n$ lobes is formed, the angular widths of the lobes being inversely proportional to n. By neglecting unwanted lobes, a $\cos n\theta$ diagram of the above kind may be made to define a kind of polar diagram which is of importance in laying down envelope spiral-phase fields.

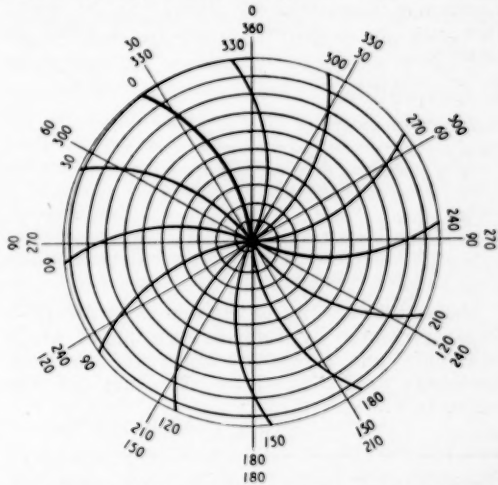


Fig. 2. *Archimedes spirals defining constant phase contours of two oppositely-directed spiral-phase fields of widely different pitch.*

For the present purpose a $\cos n\theta$ polar diagram will be considered to have only a single lobe radiating into an angle in the plane of θ lying within limits $\pm \pi/2n$ each side of the direction of maximum radiation. The radius vector defining the magnitude of radiation in the direction of maximum radiation will be called the leading radius. The plane of θ is usually horizontal.

If four aerial arrays are arranged with their leading radii as shown in Fig. 3 and if the carrier currents fed to each array are modulated progressively, as defined below, with a sinusoidal modulating frequency, then a field is set up in which the contours of constant referred phase are true Archimedes spirals.

3. Archimedes Spirals

The term progressive modulation here refers to the condition in which the phase of the

modulation advances from array to array by an angle equal to n times the difference in bearing between arrays; in the case above the value of n is unity. More generally, if a symmetrical system of m arrays having cos $n\theta$ polar diagrams, where $n = m/4$ and m is any multiple of 4, is set up, with differences in bearing between adjacent leading radii all equal to $2\pi/m$ radians, and if the carrier currents in the arrays are modulated progressively, a multiple-envelope spiral-phase field is set up in which the radius phase is proportional to n times the bearing of the radius contour defining the phase.

Within the limits of the above specification a series of assemblies of arrays can be built. This may be described in analytical terms as follows:—

4. Conventions

m = number arrays
β = 360 °$/m$ = space angle between arrays
α = phase angle between arrays = $n\beta$
θ = angle in azimuth
ϕ = radius phase
$n = \phi/\theta$ = coefficient of θ in the expression cos $n\theta$ defining the form of the polar diagram of each array of the assembly.

The series of series of assemblies of arrays is then represented by Table I, each line of which represents one series of which the different assemblies are obtained by allocating different values to n.

TABLE I

Series No.	a	m	β
2 : n	90°	$4n$	360°$/m$
3 : n	60°	$6n$	,,
4 : n	45°	$8n$	,,
5 : n	36°	$10n$	,,
6 : n	30°	$12n$	,,

In any assembly which is capable of laying down an Archimedes field, the number of polar diagrams which overlap on any one bearing is equal to $m/2n$, which is an integer. Such an assembly will be called an Archimedes assembly. Any member of the above family in which $m/2n$ is not an integer is a mongrel and exhibits the feature that the number of overlapping polar diagrams on any bearing varies from sector to sector. This means that the lines of constant referred phase are not Archimedes spirals. In the limiting case where certain sectors are covered by only one polar diagram the value of $d\phi/d\theta$ in these sectors is zero: the spiral lines of constant referred phase are degraded to arcs of circles.

It is evident that each series is characterized by the number of polar diagrams which overlap on any given bearing: this number constitutes the first part of the series number given in the first column of Table I. Each series is also characterized by the values of n, α and m.

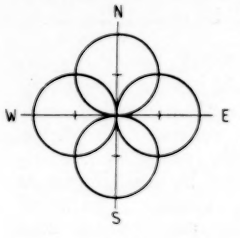


Fig. 3. (*Left*). *Relative orientation of polar diagrams of four aerial arrays.*

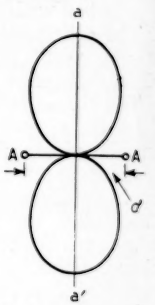


Fig. 4. (*Right*). *Polar diagram of two vertical aerials with half-wave spacing driven in phase.*

Each member array assembly of each series is then characterized by the same quantities and may be identified by two numbers, the first giving the number of overlaps and the second giving the value of n: e.g., 2 : 1, 2 : 2, etc., 3 : 1, 3 : 2 etc. In general only the assemblies with few overlaps are of interest, and up to now only an equivalent of the 2 : 1 array has been used.

5. Practical Approximation to 3 : 2 Assembly

Fig. 5 shows an assembly of three of the arrays shown in Fig. 4, which approximates to the 3 : 2 assembly. It is not exactly a 3 : 2 assembly because the polar diagrams are only approximations to cos 2θ diagrams. If all arrays are driven in phase at the carrier frequency and such an assembly is progressively modulated, the radius phase $\phi = 2\theta$ with a deviation of about $\pm 3°$, the actual value of ϕ being given by:

$$\phi = \tan^{-1} \frac{1.732 \sin\left[\frac{0.866\pi d}{\lambda}\cos\theta\right]\sin\left[\frac{\pi d}{2\lambda}\sin\theta\right]}{\cos\left[\frac{\pi d}{\lambda}\sin\theta\right] - \cos\left[\frac{0.866\pi d}{\lambda}\right]\cos\theta\cos\left[\frac{\pi d}{2\lambda}\sin\theta\right]}$$

where d is the spacing between aerials of an array (i.e., as in Fig. 4) and λ is the wavelength.

Since the relation between ϕ and θ is known, it is evidently a comparatively simple matter to calibrate any indicating instrument to read directly in azimuth.

The assembly of Fig. 5 (half-wave spacing assumed) has the advantage that the form of field is easily determinable, and depends on the fundamental polar diagrams characteristic of a simple arrangement: there is no question of faking polar diagrams to approximate to a $\cos n\theta$ diagram. It has the disadvantage of ambiguity in angular location as compared with a 2 : 1 assembly in which $\phi = \theta$, since one half of the field of radius phase contours is duplicated in the other half of the field. Provided means can be introduced to offset such ambiguity (e.g., by means of a compass and crude d.f. gear) such an assembly has an advantage of two to one, over a 2 : 1 assembly, in accuracy of angular location.

It will now be clear that the radius phase is of particular significance and that in Archimedes fields the radius phase is proportional to azimuth angle measured from the centre of the field. In spiral-phase fields other than Archimedes fields, while the radius phase is not proportional to azimuth angle, it may be a known function of azimuth angle, as for instance in the case of the assembly of Fig. 5. It follows that, while Archimedes fields appear to have one desirable characteristic, practical considerations may make the use of other types of spiral phase fields more useful.

An interesting field which has not yet been explored is the use of non-sinusoidal envelopes.

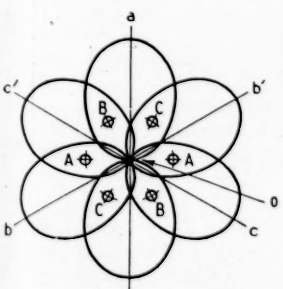


Fig. 5. *Relative orientation of polar diagrams of three aerial arrays.*

6. Applications

In all applications, it will be assumed that sinusoidally-modulated fields or their equivalent are used, and that all references to phase refer to envelope phase, unless otherwise specified. The contours defining phase relations in all fields will be assumed to lie in a horizontal plane.

Three main applications appear feasible.

(i) *Radar Range.* By setting up concentric spiral-phase and circular-phase fields, and equipping craft with means for selective reception of these fields, or the envelopes of these fields, a direct reading of radius phase can be given on a simple form of phase-difference indicator carried in the craft.

This evidently enables the craft to know the bearing of the centre of the field from the aircraft.

(ii) *Lighthouse or Talking Beacon.* By setting up a single envelope spiral-phase field, a craft may be equipped so that a gate in a radio receiver in the craft, on an independent speech or telegraph channel, opens at a time determined by the referred-phase difference, and remains open for any required short period of time. This enables a time selection system of communication to be operated in such a way that in effect continuous information may be sent out on each bearing characteristic of that bearing. This is an extremely important application because existing talking lighthouses only send out such information at intervals which are so long that an aircraft may pass the locality of the beacon without receiving an indication of bearing.

(iii) *Substitute for Narrow-beam Arrays.* (a) *Secondary Radar.* By setting up a single envelope-phase field and a concentric-envelope circular-phase field the radius phase can be used to define a single radius (e.g., where the radius phase is zero) on which coincidence gates in a craft will enable a signal to be received to initiate a return pulse sent out by a responder in the craft. If the envelope frequencies of circular- and spiral-phase fields differ, the defined radius will rotate at the difference frequency. This gives rise to an interrogating pencil of width dependent on the envelope frequencies and the minimum practical gate widths and pulse widths.

(b) *Primary Radar.* This operates on the same principle as the secondary radar except that the coincidence gates must now be located at the interrogator, and the interrogator must be provided with some means of differentiating between outgoing and incoming pulses.

Since this article was first written the question of primary radar has been investigated further, and while an apparently feasible system has been evolved on paper it is rather complex. Further, the range obtainable on primary radar with an all round looking aerial system is small.

7. Radar Range

It has already been indicated that a craft can be fitted with such equipment as will enable it to measure radius phase and that this can be used as an indication of bearing.

An American equipment making use of a 2 : 1 assembly laying down a single spiral-phase field, in conjunction with a concentric circular-phase field has been described.*

The application of spiral-phase fields to radar

*Report No. T.S. ELC/SP/2 issued by the Electronics Sub-division of Advisory Group, Air Navigation, Air Technical Service Command.

range is described in British Patent Application No. 13278/46.

8. 3:2 System

As the value of n is increased, an increase in angular accuracy of location is obtained at the expense of an increase in ambiguity of position. The value of n in a 3 : 2 system is 2, so that the ambiguity is only twofold; one half of the field repeats the other half, and it is a comparatively simple matter to provide auxiliary aids to resolve the ambiguity. The increase in angular location with increase of n follows directly from the fact that $d\phi/d\theta = \alpha/\beta = n$.

Amplifying the last paragraph slightly, it has already been pointed out that, in an Archimedes assembly of arrays laying down cos $n\theta$ diagrams, the phase rotation per circuit of the field centre is $2n\pi$. This means that in a horizontal field the change of radius phase is n times the change of azimuth bearing of the observation point from the field centre; hence the accuracy of angular location is increased n times.

A type of arrangement which approximates to an ideal 3 : 2 system will now be described, making use of the assembly in Fig. 5. As has already been explained, with this arrangement the radius phase is not exactly proportional to azimuth angle and correction has to be made for the deviation of radius phase from the azimuth angle.

With this system it is possible to make use of a single carrier frequency only. This is because the phase of the carrier field due to the aerials in the hexagon (see Fig. 5) is everywhere the same as that of an aerial located at the centre of the hexagon and driven in phase with the aerials in the hexagon. Hence the reference aerial at the centre of the hexagon may be driven with a carrier of the same frequency and in the same phase as the carrier driving the aerials in the hexagon. Also, if this central reference aerial is a plain vertical aerial, the plane of polarization of the field radiated by the central aerial is everywhere the same as that of the field radiated by the aerials in the hexagon. The net result is that the combination of the fields from the aerials in the hexagon and from the central aerial is always directly additive: no distortion is introduced by the addition of quantities in quadrature in time or in space.

For convenience in explanation a distinction will be made between the *indicating* waves radiated by the aerials in the hexagon and the *reference* waves radiated by the reference aerials at the centre of the hexagon. It is evident that, since the only information peculiar to either set of waves is in the sidebands, it is immaterial whether the unmodulated carrier component is radiated from the central aerial or simultaneously in equal amplitude and phase, from all aerials in the hexagon, or from both sets of aerials together.

In the present description, for simplicity, it will be assumed that the reference waves are obtained by modulating the carrier radiated from the central aerial with a frequency $\frac{1}{2}f$ and that the indicating waves are obtained by modulating the carrier radiated from the hexagon aerials with a frequency f, the frequency f being obtained from the frequency $\frac{1}{2}f$ by means of a frequency-doubler so that the two modulating frequencies are mutually locked.

A receiver in a craft situated in the two fields, when tuned to the carrier frequency, will deliver at the output of its detector a frequency $\frac{1}{2}f$, of phase depending only on the distance of the craft from the central aerial, and a frequency f, of phase dependent on the distance and bearing of the craft from the central aerial. These two frequencies are then separated by means of filters and the frequency $\frac{1}{2}f$ is then doubled by means of a frequency-doubler and the phase of the two waves of frequency f is then compared by means of any suitable form of phase-difference indicating meter. The indication of this meter is then equal to the radius phase plus any extra phase shifts which may exist in the system. By proper precautions the extra phase shifts may be made constant, and may be eliminated by suitable calibration of the indicating phase meter.

9. Lighthouse or Talking Beacon

A form of beacon using spiral phase fields is described in British Patent Application No. 13279/46.

This may make use of a single spiral phase field which will be assumed to be of the Archimedes type so that radius phase is proportional to azimuth and by suitable orientation can be made equal to azimuth. Departures from this condition can however be compensated.

It is evident that the phase of the field can be made to control the time at which a gate in a receiver in a craft at any point P in the field opens. Therefore a short-duration signal sent out by an independent transmitter at the beacon location, at a specified time in the cycle of rotation of the spiral-phase field, may be selected by the opening of the gate, at a specified time for a specified duration, to the exclusion of other short pulses sent out at other times.

Consequently a time distributor or selector can be provided at the transmitter to open a number of gates in turn, each for a short period, each gate connecting one communication channel to a common transmitting path. Each observation point will then receive communica-

tions from a single communication channel only. In order to achieve this it is necessary for each gate to open for the same length of time as, and in displaced isochronism (as defined below) with, each of a number of gates under the control of the radius phase at each of a number of observation points remote from the transmitter and on bearings from the transmitter which are distinctively different.

Bearings are considered to be distinctively different when, during the period of opening of one gate at the observation point, one gate and one gate only is open at the transmitting station.

The term ' displaced isochronism ' defines the condition in which the time of opening of a gate at P is delayed with regard to the opening of the corresponding gate at the transmitter by a time equal to the time of propagation of the waves from the transmitter to P. It is evident that since the opening of the gate at P is under control of a wave originated at the transmitter, if the condition of displaced isochronism exists at one range from the transmitter to P, it exists for all ranges.

For simplicity a variation will be described in which the radius phase defined by the beacon is equal to the bearing of P from O, the centre of the transmitter system. Continuing the description on this basis, it is immediately apparent that the facility is offered of arranging that one communication channel is effectively in communication with each discrete sector (a discrete sector will here be defined as a sector bounded by such bearings that a gate at P controlled as described is open only during the period of opening of one gate at the transmitting station delayed by the time of propagation). Evidently each communication channel may be arranged to send information characteristic of the bearing of the centre of the sector with which it is effectively in communication.

Any one period in which a communication channel at the beacon site is in communication with a particular sector is called a sector period. The minimum element of intelligence which it is convenient to convey as an integral whole, is called a baud.

Two modes of use of the proposed system are envisaged. In the first of these an intelligible item of information such as a Morse letter or number, or assembly of letters or words, or a word or phrase of telephonic speech, is conveyed during each sector period. In the second of these, during each sector period there is conveyed only one baud. The arrangement is however in no way restricted to these limiting conditions; during each sector there may be conveyed any intermediate number of bauds, or any fraction of a discrete intelligible phrase or assembly of symbols.

It will be evident that the first method of use is subject to the disadvantage that the frequency of repetition of each item of information on any one sector depends on the amount of information being disseminated in other sectors. If, for instance, it is required to transmit two letters with one repetition on each of 300 sectors, and the rate of signalling is 150 letters per minute, which is a fairly high rate for beacon work, the bearing information will be sent out to each sector once every eight minutes approximately. During this time an aircraft flying at the moderate speed of 300 miles an hour will have travelled 40 miles. It is, therefore, very desirable that the information should be sent out continuously in each sector.

The second method of use, where only one baud is sent out per sector period, enables information to be sent out effectively continuously to all sectors simultaneously. Assume that letters are sent out by Morse code using the first 300 pairs of letters in the several series AA, AB, AC, etc., BA, BB, BC, etc., etc., each pair being continuously reiterated in its appropriate sector. Each sector has an angular width of 1° 12′. Assume also that the letters are sent out by Morse code of standard pattern in which the length of a dot is defined as one baud, a dash is three bauds, a space between dash and dot or dot and dash in the same letter is one baud, space between letters in the same group is 3 bauds, space between groups is 5 bauds, The first group in the series is then AA and the last group is LN and the total number of bauds for one cycle of sectors, involving the transmission of 300 pairs of letters is about 7140. Since, however, the cycle is not complete until the last baud of the last letter has been sent and since some pairs contain more bauds than others it is more useful to think in terms of the total number of baud intervals, whether occupied or unoccupied, and this is obtained by multiplying the number of bauds in the pair of letters containing the most bauds by 300 and adding 1500 to the bauds in 300 inter-pair spaces of 5 bauds each. The pairs JQ and JY contain 29 bauds each so that the total effective requirement of baud intervals per complete cycle of events is $29 \times 300 + 1500 = 10{,}200$. It may be simpler to regard the space between pairs as a necessary adjunct of a pair so that the total number of bauds is $(29+5) \times 300 = 10{,}200$.

In one complete cycle of events the rotating field therefore makes 34 (i.e., 29 + 5) revolutions, and sends out 300 pairs of letters, each pair being sent out once into its appropriate sector only.

Suppose for instance that it is required to secure an average transmitting speed of 60 letters per minute on each channel; i.e., in each sector.

This corresponds to a total number of letters transmitted equal to 60×300 per minute or 150 pairs per second It could therefore be achieved by a system in which the rotating field made 17 revolutions per second and sent out 10,200 bauds conveying the 300 pairs of letters in a complete cycle of events lasting two seconds. Every two seconds, therefore, the pair of letters appropriate to each sector would be transmitted to each sector at a rate corresponding to a speed in letters per minute of 60. It may be useful to note here that at the generally recognised rate of 5 letters to a word, 60 letters per minute corresponds to the modest Morse speed of 12 words a minute which is achieved with very little practice.

The modulating frequency necessary to rotate the field, and give rise to a spiral-phase field, is only 17 c/s, a frequency at which few difficulties are likely to be experienced.

It is evident that by using bauds of sufficiently short duration speech channels may be set up from the beacon to each of a small number of selected sectors. Each channel is then constituted as a pulse-modulated channel.

10. Substitute for Narrow-beam Arrays

Narrow-beam arrays are subject to undesirable effects due to side lobes which sometimes give rise to spurious indications and errors. They make it necessary either to use very high frequencies or else to use a very large array.

It is evident that for purposes where the scanning of a region is required it should be possible to make use of the rotation of a spiral phase field. It appears that it is possible to produce by this means a narrow exploring pencil, radiating from and rotating round any required location, without the use of mechanically rotating arrays, and without the use of arrays having narrow polar diagrams.

To this end a system is set up producing a single Archimedes envelope field concentric with an envelope circular phase field.

Definition:—The point in time at which any sinusoidal varying quantity passes through zero in a direction of increasing positive amplitude will be called the *zero plus* point.

Secondary Radar.—If the envelope of the spiral-phase field is of the same frequency as that of the circular-phase field, the locus in space, defining the location where the zero plus points of each of the envelopes of each field observed separately occur simultaneously, is a fixed radius vector with one end located at the centre of the system laying down the spiral phase-field.

This radius vector is called the simultaneous radius.

If the envelope field frequencies differ by d cycles per second, then the simultaneous radius rotates making d revolutions per second, in the same sense as the spiral-phase field if the envelope of that field is of higher frequency, otherwise at d cycles per second in the opposite direction.

If now a responder is provided which only makes response when situated within a narrow sector each side of the immediate vicinity of the simultaneous radius, it is evident that a means is available for locating the position of any craft carrying such a responder, within the limits of error defined by the minimum dimensions which it is possible to realise for the said sector.

In order to enable the two envelope frequencies corresponding respectively to the two envelopes to be separated at the receiver, it is necessary to arrange that each envelope field is constituted by the modulation of a separate carrier or sub-carrier. More specifically one arrangement is as follows. The carrier frequency of each transmitter connected to the arrays which are to radiate the spiral phase fields is modulated with one or two frequencies which will be referred to as sub-carriers. These sub-carriers are arranged to be so far apart in frequency that no difficulty is experienced in separating them by filters. One sub-carrier is then always modulated by the spiral-phase-envelope-field-producing (s.p.e.f.p.) frequency and used to modulate the main carriers fed to the array producing the spiral phase envelope field. The other sub-carrier is modulated with the circular-phase-envelope-field-producing (c.p.e.f.p.) frequency and used to modulate the main carrier fed to the central vertical aerial.

After normal detection at the receiver, the two sub-carriers appear, one modulated by the (s.p.e.f.p.) in phase corresponding to the angular position of the receiver with regard to the centre of the region where the array of arrays producing the spiral phase envelope fields is situated, the other modulated by the (c.p.e.f.p.) frequency.

One of these frequencies is arranged to open a gate at the zero plus point in its cycle ; the other frequency is arranged to initiate, at the zero plus point in its cycle, a pulse of duration equal to the period for which the gate is opened by the first frequency. It will be immediately apparent that, when the two zero plus points of the received envelopes are substantially coincident, which is when the receiver is on or near the simultaneous radius, the transmitter is modulated by a pulse-recurrence frequency equal to that of the c.p.e.f.p. frequency. Evidently, if the gates are of zero width in time the angular width of the sector each side of the simultaneous radius over which transmission of a pulse will occur will be zero. In practice the gates and, therefore, the sector will be finite in width.

It will be further evident that the accuracy of location provided by this system can be increased at the cost of ambiguity by using any form of envelope spiral-phase field in which the change of phase shift with change of angular position is greater than unity.

When, however, it is required to scan in a limited sector, provided the sector does not subtend an angle greater than $360°/n$ no ambiguity need arise.

It has not been previously pointed out that in order to produce a spiral phase field in a limited sector it is only necessary to provide and drive the aerials of the system laying down a spiral-phase field system, which are so orientated and have such polar diagrams that they radiate energy into that sector.

For instance, in the case of a single-phase Archimedean envelope spiral-phase system in which four aerials with unilateral cos θ diagrams, orientated respectively North, South, East and West, are driven in progressive quadrature it is possible to set up a spiral-phase field in any of the quadrants NE, SE, SW, or NW, by using only the two aerials which radiate into that quadrant.

11. Theory

To establish that m arrays having cos $n\theta$ diagrams, with $m = 4n$, when progressively modulated, will lay down lines of constant referred phase which are Archimedes spirals and that in such a field the radius phase is proportional to bearing.

f = frequency

$\omega = 2\pi f$

r = radial distance (of point P at which field is observed) from centre of assembly of arrays.

c = velocity of light

$\phi_c = \frac{\omega r}{c}$ = phase shift retardation due to spatial displacement by distance r.

θ = bearing (or azimuth angle) taken from the centre of the assembly of arrays.

ϕ_r = referred phase difference

ϕ_d = radius phase $= \phi_r - \phi_c$

α = phase angle between carriers or modulation frequencies of arrays.

$\beta = \frac{360°}{m}$ = space angle between arrays

$n = \frac{\alpha}{\beta} = \frac{\phi_d}{\theta}$ = coefficient of θ in the function cos $n\theta$ defining polar diagram of array

Note that $\alpha = n\beta$

m = number of arrays

M/ϕ_r = field at point P distance r from centre of assembly on bearing θ

e = maximum contribution of an array (i.e., on leading radius) to field at distance r from centre of assembly.

Then

$$M/\phi r = e/\underline{-\phi_c} \cos n\theta + e/\underline{\alpha - \phi_c} \cos n(\theta-\beta) + e/\underline{2\alpha - \phi_c} \cos n(\theta - 2\beta) + e/\underline{2\alpha - \phi_c} \cos n(\theta - 3\beta)$$

The number of terms in the series determining M/ϕ_r is equal to the number of polar diagrams which overlap on any one bearing; i.e., to $m/2n$.

By taking any required number of terms in the above series, the numerical values characteristic of an assembly, with the corresponding number of overlaps, can be found for both Archimedes and mongrel (non-Archimedes) assemblies.

Two-term Arrays.

$$M/\underline{\phi_r} = e/\underline{-\phi_c} \cos n\theta + e/\underline{\alpha - \phi_c} \cos n(\theta - \beta)$$

Making use of the transformation,

$$\cos n\theta = \tfrac{1}{2}/\underline{n\theta} + \tfrac{1}{2}\overline{\backslash n\theta}$$

$$M/\underline{\phi_r} = \frac{e}{2}\overline{\backslash\phi_c}\ (/\underline{n\theta} + \overline{\backslash n\theta} + /\underline{n(\theta-\beta) + \alpha} + \overline{\backslash n(\theta-\beta) - \alpha})$$

Since $\alpha = n\beta$, the first and third terms add up to $e/\underline{n\theta - \phi_c}$, which represents a phase angle proportional to n times the azimuth angle minus the phase retardation ϕ_c: in other words $\phi_r = n\theta - \phi_c$ which is the condition defining an Archimedes spiral-phase field.

It remains to establish conditions which make the remaining terms zero; that is:

$$\overline{\backslash n\theta} = -\overline{\backslash n(\theta-\beta) - \alpha}$$

This evidently occurs when

$$n\beta + \alpha = \pm 180°, \text{ and since } n\beta = \alpha$$
$$\alpha = 90°$$

Evidently $m = \frac{360°}{\beta} = \frac{360°}{\alpha/n} = 4n$

Summary: $\alpha = 90°$, $m = 4n$, $\beta = \frac{360°}{m}$

This summary evidently defines the first of the series of assemblies listed in Table I.

Three-term Assemblies

$$M/\underline{\phi_r} = \frac{e}{2}\overline{\backslash\phi_c}\ (/\underline{n\theta} + \overline{\backslash n\theta} + /\underline{n(\theta-\beta) + \alpha} + \overline{\backslash n(\theta-\beta) - \alpha} + /\underline{n(\theta-2\beta) + 2\alpha} + \overline{\backslash n(\theta-2\beta) + 2\alpha}$$

Since $\alpha = n\beta$ the three odd terms add up to $\frac{3}{2}\, e/\underline{n\theta - \phi_c}$ which defines an Archimedes field.

The three even terms are zero when they

constitute three equal vectors at 120°, that is when $n\beta + \alpha = \pm 120°$

so that $\alpha = 60°$

and $m = \frac{360°}{\alpha} = \frac{360°}{\alpha/n} = 6n$

Summary: $\alpha = 60°$, $m = 6n$, $\beta = \frac{360°}{m}$

This evidently defines the second series of assemblies listed in Table I.

The characteristics of the other assemblies can evidently be found by taking four, five, six, etc., terms.

To show that the condition that $M/\phi_r = e/n\theta - \phi_c$ defines an Archimedes field.

Evidently $\phi_r - n\theta - \phi_c = n\theta - \frac{\omega r}{c}$

$\therefore r = -\frac{c\phi_r}{\omega} + \frac{c}{\omega} n\theta$

For any constant value of ϕ_r the radius r is defined by a quantity $\frac{c}{\omega} n\theta$ which is proportional to azimuth angle plus or minus a constant quantity $\frac{c}{\omega} \phi_r$.

The value of r is therefore proportional to azimuth angle, and the curve so defined is an Archimedes spiral

To show that in an Archimedes field the radius phase is proportional to bearing.

In an Archimedes field, the referred phase difference $= \phi_r = n\theta - \phi_c$.

In a circular-phase field the referred phase difference $= -\phi_c$.

The radius phase is equal to the phase difference between an Archimedes field and a concentric circular-phase field and is therefore equal to the difference between the referred phase of these fields ; i.e.,

Radius Phase $= \phi_r + \phi_c = n\theta$.

Example of a Theoretical Examination of a Simple Practical Assembly.

In practice arrays with cos $n\theta$ diagrams are not always readily available and it is necessary to choose some simple arrangement which conforms approximately to an assembly of arrays with cos $n\theta$ diagrams. One such arrangement has been indicated in Fig. 5.

Let the field due to any one aerial at any point P distance r from any one aerial be $E \sin \omega t$. The phase retardation has been omitted because it cancels out in the analysis below.

Assume a modulation depth (of each polar diagram.) m by a sinusoidal modulation of (angular) frequency M, and a progressive phase shift of 120° in phase from array to array.

The field at P due to array AA in Fig. 5 may then be represented by

$$(1 + m \sin Mt) \times \left[E \sin \left(\omega t + \frac{2\pi d/2}{\lambda} \sin \theta \right) + E \sin \left(\omega t - \frac{2\pi d/2}{\lambda} \sin \theta \right) \right]$$

where d is the distance between aerials and λ is the wavelength.

This is equal to

$$2E\,(1 + m \sin Mt) \cos \left(\frac{\pi d}{\lambda} \sin \theta \right) \sin \omega t$$

Since the modulation advances 120° in phase from array to array, and the azimuth angular displacement from array to array is 60°, the fields due to the other two arrays are :

$$2E(1 + m \sin \overline{Mt + 120°}) \cos \left(\frac{\pi d}{\lambda} \sin \overline{\theta - 60°} \right) \sin \omega t$$

and

$$2E(1 + m \sin \overline{Mt + 240°}) \cos \left(\frac{\pi d}{\lambda} \sin \overline{\theta - 120°} \right) \sin \omega t$$

Now let the amplitude and phase of the envelope of the total field at P with regard to the envelope of array AA be represented by A/ϕ_r.

Then adding the envelopes of the three fields and resolving in time along axes parallel and normal to the vector defining the modulation phase of array AA :

$$\begin{aligned} A/\phi_r = {} & 2mE \cos \frac{\pi d}{\lambda} \sin \theta - mE \cos \frac{\pi d}{\lambda} \sin \overline{\theta - 60°} \\ & - mE \cos \left(\frac{\pi d}{\lambda} \sin \overline{\theta - 120°} \right) \\ & + j\ 1.732\ mE \cos \left(\frac{\pi d}{\lambda} \sin \overline{\theta - 60°} \right) \\ & - j\ 1.732\ mE \cos \left(\frac{\pi d}{\lambda} \sin \overline{\theta - 120°} \right) \\ = {} & 2mE \cos \frac{\pi d}{\lambda} \sin \theta \\ & - 2mE \cos \left\{ \frac{\pi d}{\lambda} (\sin \overline{\theta - 60°} + \sin \overline{\theta - 120°}) \right\} \\ & \times \cos \left\{ \frac{\pi d}{\lambda} (\sin \overline{\theta - 60°} - \sin \overline{\theta - 120°}) \right\} \\ & + j\ 3.464\ mE \sin \left(\frac{\pi d}{\lambda} \sin \overline{\theta - 60°} \sin \overline{\theta - 120°} \right) \\ & \times \sin \left\{ \frac{\pi d}{2\lambda} (\sin \overline{\theta - 60°} - \sin \overline{\theta - 120°}) \right\} \\ = {} & 2nE \cos \frac{\pi d}{\lambda} \sin \theta \end{aligned}$$

$$-2mE\cos\left(\frac{\pi d}{\lambda}\sin\overline{\theta-90}^{\circ}\cos 30^{\circ}\right)$$
$$\times\cos\left(\frac{\pi d}{\lambda}\cos\overline{\theta-90}^{\circ}\sin 30^{\circ}\right)$$
$$+j\,3.46\,mE\sin\left(\frac{\pi d}{\lambda}\sin\overline{\theta-90}^{\circ}\cos 30^{\circ}\right)$$
$$\times\sin\left(\frac{\pi d}{\lambda}\cos\overline{\theta-90}^{\circ}\sin 30^{\circ}\right)$$
$$=2mE\cos\frac{\pi d}{\lambda}\sin\theta$$
$$-2mE\cos\left(\frac{0.886\,\pi d}{\lambda}\cos\theta\right)\cos\left(\frac{\pi d}{2\lambda}\sin\theta\right)$$
$$+j\,3.464\,mE\sin\left(\frac{0.866\,\pi d}{\lambda}\cos\theta\right)\sin\left(\frac{\pi d}{2\lambda}\sin\theta\right)$$

$$\therefore\ \phi_r=\tan^{-1}\left[\frac{1.732\sin\left(\frac{0.866\,\pi d}{\lambda}\cos\theta\right)\sin\left(\frac{\pi d}{2\lambda}\sin\theta\right)}{\cos\left(\frac{\pi d}{\lambda}\sin\theta\right)-\cos\left(\frac{0.866\,\pi d}{\lambda}\cos\theta\right)\cos\left(\frac{\pi d}{2\lambda}\sin\theta\right)}\right]$$

The relation between ϕ and θ, as calculated by slide rule, appears in Table II. It is evident that the deviation of ϕ from 2θ is not large, although the accuracy of calculation is not sufficient for the deviation to be specified accurately. The maximum deviation is however not more than a few degrees, probably about 3 degrees, and as such can easily be corrected in any system where the value of ϕ is required to indicate the value of 2θ.

TABLE II

θ	ϕ	θ	ϕ	θ	ϕ
0	0°	70	137°0′	140	277°17′
10	17°16′	80	157°0′	150	300°0′
20	37°15′	90	180°0′	160	322°45′
30	60°	100	203°0′	170	342°44′
40	82°40′	110	223°4′	180	360°
50	102°40′	120	240°12′		
60	120°	130	257°22′		

Appendix

C.A.A. V.H.F. Omnidirectional Range

This is the first system to make use of a spiral-phase field for navigational purposes; it makes use of a 2 : 1 assembly laying down a single spiral-phase field. The credit for its development belongs to American engineers, and it has been recommended by P.I.C.A.O. as a short-range navigational aid.

It is stated to provide bearing indications with an accuracy better than 5°, operating on a wavelength of 2.4 metres.

American Standard Abbreviations and Symbols

THE American Institute of Radio Engineers has recently published a 21-page report (price 75 cents) entitled "Standards on Abbreviations, Graphical Symbols, Letter Symbols, and Mathematical Signs," embodying "the Institute's official recommendations to its members, the industry at large, and the armed services." Although, of course, there is much agreement between the American and the British Standards there are many differences which make one wish that such things could be discussed internationally. It should be pointed out that the publication is not issued by the American Standards Association but by one Institute, and, although mention is made of co-operation with other bodies, it is possible that the American Institute of Electrical Engineers may not be in complete agreement with all the recommendations of the Radio Engineers. Some differences between British and American symbols are to be expected since the anode of a valve in one country is the plate of a tube in the other, but there are many quite unnecessary and surprising differences. In the decimal prefixes the only difference is in the abbreviation of kilo for which they give both k and K; both are undoubtedly largely used although, as we recently pointed out, kg and km are now standard abbreviations in all European metric countries which use these units in everyday life.

One of the most surprising things is the recommendation of small letters for all the electrical units except the ohm, for which the capital Ω is retained. Kilowatts are therefore either kw or Kw but not kW, microfarads are μf and millihenrys mh. For cycles per second they recommend cps which is further abbreviated to c or C in combinations; the latter is a strange suggestion. Whereas we write kc/s for kilocycles per second, the American radio engineer can write kc or kC or Kc or KC, always leaving the unit of time to the imagination. No misunderstanding is likely to arise from the omission of the time as the 'per second' is understood, and c could be regarded as an abbreviation, not of cycles, but of cycles per second.

The symbol for dielectric constant or capacitivity —permittivity is not mentioned—is ϵ whereas

the B.S.I. recommends κ; ϵ is certainly very widely used even in this country by physicists. Electromotive force and potential difference appear to be lumped together, for in one place we read E, e, V, v electromotive force, potential difference, voltage; in another place "V, v, E, e potential difference, electromotive force, voltage" and yet in a third place "V, v, E, e, voltage, electromotive force, potential difference." Electric field strength is represented by **E**—whereas the B.S.I. employs $\mathfrak{E}$—magnetic field strength by **H**, and magnetomotive force by a capital script $\mathcal{F}$.

Eleven pages are devoted to graphical symbols for circuit elements, coaxial and waveguide elements and valves. This is a very big subject and we would only refer to the fact that two wires crossing without connection are represented by + and not by ⊥; if the wires are connected they are represented by either ⊣⊢ or ⊥. This is a point that has recently been discussed by a committee of the B.S.I. and the consensus of opinion was in agreement with these recommendations. Although the older method is clearer and more fool-proof it is quite impracticable when there are a great number of crossing wires as in telephone-switchboard diagrams, and the question then arises as to whether it is not better to be consistent and use the straight cross-over even in the simplest cases. Some people will undoubtedly prefer to use the old-fashioned "hop over" for unconnected crossings in simple diagrams of connections.

G. W. O. H.

CORRESPONDENCE

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Ignition Interference

Sir,—I am writing to draw attention to the phenomenon of multiple-pulse interference (see Eaglesfield, *Wireless Engineer* 1946, p. 265) the importance of which is, in my view, greatly underestimated in the majority of papers on the problem of ignition interference. Thus Nethercot (*Wireless World*, October 1947) does not deal with the phenomenon at all, while Pressey and Ashwell (*Wireless Engineer*, January 1949) argue that since, on their measurements, the rest of the train is usually some 20 db down on the initial pulse, the question is not important.

Some ten years ago I observed the empirical fact that interference under actual television listening conditions from some cars was much worse than that from a laboratory generator, even when the input from the latter was very large, and I followed this up in the course of experiments upon noise suppressors. With the aid of such suppressors it was found quite easy to reduce noise from the laboratory generator (which generated single sparks spaced by 10 to 50 milliseconds) to near inaudibility for any reasonable input. On some cars, or on a special laboratory generator producing multiple sparking, however, quite loud interference still occured.

The reason for this is not far to seek: in the outer half of the television service area, where the problem of interference is most severe, the receiver gain is set to receive signals of 0.1 to 0.5 mV/m. From Pressey and Ashwell's paper, the interference 'peak field' will exceed 10 mV/m at 40ft for an aerial 30-ft high. The receiver will thus be overloaded to the extent of 20 db or more by the initial pulse, which will then appear at the output to have the same amplitude as the succeeding ones. Now it can be shown (p. 103 of Radio Research Report No. 13, confirmed by tests by the writer) that the apparent loudness of the noise is proportional to the total area of the output pulse, or pulse train, provided this does not exceed some 300 μsec. A car producing a typical half-millisecond train of a hundred pulses will thus give enormously louder interference than that from a single pulse (duration at detector order of 5 μsec). This difference in loudness is more or less independent of the integrating action of 'top-cut' circuits in the a.f. portion of the set since these cannot alter the total area, but the difference will be increased rather than otherwise by the action of a.f. noise suppressors, most of which tend (a) to equalize the amplitudes of the pulses in a train by limiting, or (b) to lose efficiency on long continued pulse trains.

This reasoning is supported by listening tests using an ordinary television receiver, incorporating a noise suppressor, on an aerial 50ft from a main road. Those cars which gave 'serious' (or 'loud') interference showed pulse trains which were relatively long (50 to 1,000 μsec), and more or less uniform in height, when examined at the output of the noise limiter with an oscilloscope arranged to synchronize off the first pulse in each train.

In order to reduce interference with television sound reception to a level where it is not in practice serious, I conclude that it is quite as important to reduce the number of pulses in the train as to reduce the field radiated. Similar conclusions probably apply to vision reception under practical conditions, though it is somewhat more difficult to find a measure for subjective interference in this case.

I suggest further that the reasons for multiple pulse generation should be seriously investigated, since the reduction of field obtained from the addition of the usual 5,000-ohm suppressors is not always sufficient completely to remove interference in difficult cases, and since manufacturers often object to the fitting of suppressors. From a cursory examination it appears that two separate phenomena are at times concerned: first, an electrical resonance involving the leakage inductance of the ignition coil and producing several main sparks spaced by the order of 100 μsec, and secondly some phenomenon in the spark itself causing each main spark to consist of a train of sub-sparks spaced by the order of 10 μsec. A research into this question (by ignition equipment makers, by the E.R.A., or by the National Physical Laboratory) might well throw light upon the problems of the ignition of mixtures for internal combustion engines as well as upon radio interference.

Southend-on-Sea, Essex. M. V. CALLENDAR.

NEW BOOKS

Elektronen Strahl Oszillographen

By PAUL E. KLEIN. Pp. 210 with 325 illustrations. Weidmannsche Verlagsbuchhandlung. Blumenstrasse 3, Frankfurt on Main. Price 19 Marks.

This book is the first of two volumes planned by the author on electron-beam oscilloscopes and their many-sided applications. The object of this first volume is to describe the tube itself and the auxiliary devices necessary for the practical application of the methods of measurement employing cathode-ray oscilloscopes. The second volume will deal with these practical applications. This is really the second edition of a single-volume book entitled 'Die praktische Verwendung des Elektronen Strahl Oszillographen"; instead of revising it the author decided to rewrite it completely in two volumes. The first chapter deals with the tube, the various methods of focusing and deflecting the beam, calculation of deflection, etc. Successive chapters are devoted to the production of the necessary high voltage, methods of connection to the deflecting plates, brightness control, amplifiers, time-base circuits, multiple oscillographs, cathode-ray photography, and the construction and operation of complete outfits. The bibliography contains 35 references, all German with one exception. The quality of the paper and binding and some of the photographic reproductions show the difficulties under which such books are being produced in Germany, but this is no excuse for 'Lissjous-Figue' under Fig. 281. The book is well illustrated and can be recommended to anyone wishing to study the development of the cathode-ray oscilloscope in Germany.

G. W. O. H.

Vacuum-Tube Circuits

By LAWRENCE BAKER ARGUIMBAU. Pp. 668 + vi with 574 illustrations. Chapman & Hall Ltd., 37, Essex St., London, W.C.2. Price 36s.

This book is of American origin, the author being Assistant Professor of Electrical Communications at the Massachusetts Institute of Technology. Anyone skimming the book will receive the impression that it is just another book on radio and a fairly elementary one at that.

Appearances are deceptive, however, and this is by no means an ordinary book. Most readers will receive a shock when they reach p. 26 and meet the first set of problems. Up to then everything has been very easy. After an opening chapter devoted to a very simple explanation of the processes used in communications, a start is made with 'Diodes and Rectifiers.' About 5½ pages are given to emission and diode characteristics and then 2½ pages to the diode as a peak voltmeter. It is to be emphasized that the treatment is quite elementary and largely descriptive; two equations only appear so far and neither refers to the voltmeter usage of the diode.

The problems then come and refer to a simple diode circuit having a load resistance R and a reservoir capacitance C. The first is—"Show that when $T \ll CR$ in Fig. 8 the average condenser voltage is given approximately by $E_c = E_{peak}\ (1 - T/2CR)$"; another is—"Assuming a very large value for C in Fig. 8, show that the effective impedance presented to the source is $R/2$."

The reader's first reaction will undoubtedly be that it is quite impossible to solve such problems as these with the information given in the preceding pages. When he thinks about it, however, he will realize that he needs only two things,—a good knowledge of circuit theory apart from valves and the explanation of the diode given in the book,—in order to solve them and that the solutions are, in fact, quite simple.

Further reading shows that the author adopts such shock tactics throughout. After describing the mode of operation of a circuit in quite simple language, a problem is given and worked out in full as an example; then come a number of problems to which answers are not given. These problems require for their solution a good knowledge of general theory, but they also demand that the relevant text in this book shall have been read carefully and *understood*.

The method is a very sound one and it is refreshing to find a book which differs so much from the general run of text-books. To benefit from it, the problems must be worked out and although they are not really difficult ones, some of them need quite hard thinking.

It is a fault of the book, which may mislead some, that the author does not point out the need for considerable prior knowledge of circuit theory apart from valves. This need for prior knowledge also detracts from the usefulness of the book to radio technicians who, quite often, tend to be weaker in circuit theory than in their knowledge of the general modes of operation of valve circuits. To those whose training has been largely electrical rather than radio, however, the book should be extremely useful.

The contents cover all the ordinary valve circuits, and amplifiers (class A, B and C) and oscillators are treated. Some circuit theory is included, notably in dealing with the stagger-tuned and band-pass r.f. amplifier and with the transient response of v.f. amplifiers. Most of the book requires only simple mathematics but there are patches of calculus and Fourier series which some will find sticky going.

The squegging oscillator (called a blocking oscillator) is rather poorly treated, neither is the blocking oscillator (called a multivibrator) well done. The author treats them as distinct phenomena and very inadequately. Similarly, the treatment of a magnetic-deflection circuit in television ignores the circuit capacitance and thereby makes it of little practical value. On the whole, however, the treatment is good and the book can be recommended to those who have the necessary knowledge of circuit theory but lack a familiarity with valve circuits and the application to them of their circuit theory.

W. T. C.

Telephony

By J. ATKINSON, A.M.I.E.E., Vol. I, General Principles and Manual Exchange Systems. Pp. 513 with 694 illustrations. Sir Isaac Pitman & Sons, Ltd., Kingsway, London, W.C.2. Price 35s.

This book is described on its title page as a new edition of "Telephony," Herbert & Proctor, and it bears the sub-title "A Detailed Exposition of the Telephone Exchange Systems of the British Post Office." Vol. II, which is in course of preparation, will deal with automatic exchange systems and the two volumes are together designed to cover Grades I, II and III of the new City & Guilds' Examinations in Telephone Exchange Systems and the telephony aspects of Elementary Telecommunications Practice.

Radio, Television and Electrical Repairs

Edited by ROY C. NORRIS. Pp. 448 with 429 illustrations. Odhams Press Ltd., Long Acre, London, W.C.2. Price 10s. 6d.

WIRELESS PATENTS

A Summary of Recently Accepted Specifications

The following abstracts are prepared, with the permission of the Controller of H.M. Stationery Office, from Specifications obtainable at the Patent Office, 25, Southampton Buildings, London, W.C.2, price 2/- each.

AERIALS AND AERIAL SYSTEMS

599 387.—Trailing aerial for aircraft in which the tuning is automatically controlled through the winding and unwinding gear.

Autophon A.G. Convention date (Switzerland) 24th December, 1943.

DIRECTIONAL AND NAVIGATIONAL SYSTEMS

598 609.—Pulsed signalling system for visually indicating to the pilot of a mobile craft his position, in terms of azimuth and distance, relatively to a fixed reference point.

Hazeltine Corporation (assignees of A. V. Loughren). Convention date (U.S.A.) 15th September, 1944.

598 672.—Automatic gain-control system, particularly for a radiolocation receiver in which the echo signals are presented side-by-side, or on separate indicator tubes, for comparison.

J. V. Dunworth. Application date 28th June, 1945.

598 701.—Radiolocation equipment for centimetre waves, wherein the time interval is accurately measured in terms of the number of complete cycles of an auxiliary synchronized oscillatory voltage.

J. D. Cockcroft, P. E. Pollard, B. Newsom, The British Thomson-Houston Co. Ltd. and G. S. C Lucas. Application date 30th March, 1943.

598 900.—Radio signalling system for marking-out a water-channel, or other tortuous course, by the radiation from a number of electromagnetic 'horn' aerials placed at strategic points.

Western Electric Co. Inc. Convention date (U.S.A.) 5th April, 1940.

598 948.—Radiolocation receiver wherein automatic gain-control is applied to increase gradually the sensitivity of the indicator to the more-distant objects.

Sperry Gyroscopic Co. Inc. (assignees of G. E. White). Convention date (U.S.A.) 22nd January, 1943.

598 973.—Circuit arrangement for generating the exponential voltages used for controlling the cathode-ray indicators in radiolocation.

Marconi's W.T. Co. Ltd. (assignees of M. H. Mesner). Convention date (U.S.A.) 20th August, 1942.

598 977.—Beam aerial comprising a given length of waveguide with spaced and graded apertures, which are backed by rotary reflecting discs in order to swing the beam to-and-fro.

Marconi's W.T. Co. Ltd. (assignees of H. B. de Vore). Convention date (U.S.A.) 9th June 1944.

599 087.—Direction-finding system, operating on pulsed signals and comprising a switch-controlled receiver for separately comparing the response from two aerials of different characteristics.

B. S. Smith and F. E. J. Girling. Application date 30th June, 1945.

599 114.—Radiolocation apparatus utilizing the Doppler frequency-shift, wherein a .local reference-frequency having a fixed phase-relation with the exploring pulses is derived from supersonic oscillations set up by a piezo-electric crystal along a mechanical line resonator.

A. E. Solley. Application date 13th September, 1945.

599 389.—Automatic gain-control system applied to the echo signals received in the different sectors of a rotary exploring beam, in radiolocation.

J. M. Robson. Application date 2nd May, 1945.

599 497.—Coupling bridge-network, including a phasing device, arranged to allow two aerials having different polar diagrams to be continuously interchanged, say for radiolocation.

C. H. Westcott. Application date 9th January, 1945.

599 551.—Direction finder using a rotating aerial and a visual indicator, wherein auxiliary optical means are provided for adjusting the normal alignment of the signal on the indicator.

Standard Telephones and Cables Ltd. (assignees of G. S. Burroughs). Convention date (U.S.A.) 26th January, 1944.

600 166.—Automatic gain-control system, particularly for regulating the output of a radiolocation receiver.

Sperry Gyroscope Co. Inc. Convention date (U.S.A.), 29th September, 1944.

600 327.—Directive aerial system comprising a parabolic reflector with a focal length only one-fifth that of the carrier wave, and a broad-band radiating element located at the focal point.

Marconi's W.T. Co. Ltd. (assignees of G. H. Brown and O. McD. Woodward Junr.). Convention date (U.S.A.) 22nd January, 1944.

600 433.—Aerial for ultra-short waves in which the action of an electromagnetic horn or parabolic mirror is controlled or modified by resonant slots.

H. G. Booker. Application date 31st October, 1945.

600 483.—Direction finder with a switched c.r. indicator which is fed through a common amplifying channel, wherein periodic damping is applied in order to prevent distortion.

F. Caplin and J. H. Bagley. Application date 7th June 1945.

600 494.—Radiolocation equipment in which the gain control is automatically varied between given limits by a multivibrator device, in order to minimize the effect of ground signals, 'clutter,' and the like.

R. H. A. Carter. Application date 14th September, 1945.

600 512.—Control circuit for the electrodes of a c.r. indicator tube, particularly for preventing misalignment between the line-of-sight and the beam axis, in radiolocation.

R. S. Webley. Application date 2nd May, 1945.

600 514.—Time base and marker pulse generating and triggering circuits, applied to the measurement of short time intervals, particularly in radiolocation.

A. Harrison. Application date 2nd June, 1945.

600 546.—Pulsed radiolocation equipment for use on a mobile craft, wherein provision is made to offset the effects of yawing and pitching.

Sperry Gyroscope Co. Inc. (*assignees of J. J. Caldwell, Jr. and R. F. Hays, Jr.*). *Convention date* (*U.S.A.*) *19th May*, 1942.

600 586.—Radiolocation equipment in which the echo signals from two flat exploring beams are applied to follow the course, and indicate the origin, of a shell or other projectile.

A. E. Kempton and D. R. Chick. Application date 23rd October, 1945.

600 622.—Direction finder in which signals from different aerials are fed to a cathode-ray indicator through separate amplifiying channels, the gain of which is automatically equalized by means of an auxiliary oscillator.

Soc. Francaise Radio-Electrique. Convention date (*France*), 14th February, 1942.

600 633.—Adapting a normally non-directive aerial of the 'Alford Loop' type so that its radiation pattern is made directive.

Standard Telephones and Cables Ltd. (*assignees of A. G. Kandoian*). *Convention date* (*U.S..A.*) *17th July*, 1944..

RECEIVING CIRCUITS AND APPARATUS

(*See also under Television*)

598 398.—Super-regenerative receiver, with cathode load resistance, for responding to interrogating signals and triggering an identification signal in reply.

Ferranti Ltd., M. K. Taylor and F. C. Williams. Application date 5th November, 1945.

598 399.—Coupling arrangement between the super-regenerative receiver and the triggered transmitter of an automatic 'responder' or identification unit.

Ferranti Ltd. and M. K. Taylor. Application date 5th November, 1945.

598 400.—Control arrangement for stabilizing the operation of an identification unit designed to respond automatically to interrogating signals.

Ferranti Ltd., F. C. Williams and J. R. Whitehead. Application date 5th November, 1945.

598 401.—Controlling the sensitivity of the super-regenerative receiver, and the stability of the associated transmitter, in an identification unit which responds automatically to interrogating signals.

Ferranti Ltd. and H. Wood. Application date 5th November, 1945.

598 662.—Selective coupling arrangement, say for a superheterodyne, in which an increase in coupling causes the separate peaks of the respective circuits to move further apart.

Murphy Radio Ltd., P. C. Cullen and C. C. Holmes. Application date 24th November, 1944.

598 721.—Band-pass coupling, comprising first and second transformers, wherein the secondary winding of the first is reactively coupled to the primary winding of the second, and conversely, in order to widen the response curve.

K. W. Horner. Application date 7th September, 1945.

598 807.—Super-regenerative receiver wherein the input circuit is coupled to the oscillatory circuit of the amplifier in rhythm with the quenching frequency, in order to prevent re-radiation.

"Patelhold" Patentverwertungs &c. A.G. Convention date (*Switzerland*), *11th September*, 1944.

598 868.—Receiving circuit for frequency-modulated signals, comprising a cathode-follower amplifier and a pair of diodes, arranged so as to throw very little load across the input circuit.

Marconi's W.T. Co. Ltd. (*assignees of K. Schlesinger*). *Convention date* (*U.S.A.*), *21st September*, 1944.

599 201.—Intervalve coupling for h.f. and i.f. amplifiers in which a tapped inductance is used both to provide a band-pass characteristic and to match the valve impedances.

P.R.T. Laboratories Ltd. and W. E. Benham. Application dates 11th May, 1945, *and 11th April*, 1946.

599 341.—Construction, mounting, and processing of contact detectors of the silicon type, particularly for high-frequency work.

Western Electric Co. Inc. Convention date (*U.S.A.*) *24th March*, 1944.

599 424.—Super-regenerative receiver in which a grounded-grid valve is directly coupled to the regenerative circuits and serves as a buffer to prevent re-radiation from the aerial.

Hazeltine Corporation (*assignees of B. D. Loughlin*). *Convention date* (*U.S.A.*) *1st March*, 1945.

TELEVISION CIRCUITS AND APPARATUS

FOR TRANSMISSION AND RECEPTION

598 364.—Projecting large-scale television pictures from a photographic film that is prepared from the image produced on the sensitive screen of a cathode-ray tube.

F. B. Dehn (*communicated by A. B. Dumont Laboratories Inc.*). *Application date 13th March*, 1945.

598 792.—Television receiver in which the fluorescent screen of a cathode-ray tube is of globular form, and is rotated synchronously with the scanning speed, in order to project the picture on to a viewing screen.

The General Electric Co. Ltd. and L. C. Jesty. Application date 10th September, 1945.

598 802.—Television receiver in which the electron stream scans a thin film of oil on a slowly-rotating glass disc, so as to vary its transparency to an external source of light, thus allowing the picture to be projected on to an external screen.

W. Ehrenberg. Application date 11th September, 1945.

599 134.—Colour television system in which the effective areas of the different filters are automatically interrelated so as to preserve a correct colour balance for all adjustments.

Radio Corporation of America. Convention date (*U.S.A.*) *29th November*, 1943.

TRANSMITTING CIRCUITS AND APPARATUS

(*See also under Television*)

598 375.—Coupling-device between two offset and relatively-rotatable waveguides, in which the opposite ends of a complex coaxial-line connector serve as probes.

R. G. Garfitt. Application date 7th May, 1945.

598 493.—Waveguide or transmission line composed of materials having graded values of magnetic permeability and dielectric constant, chosen to secure a predetermined mode of propagation.

Hazeltine Corporation (*assignees of H. A. Wheeler*). *Convention date* (*U.S.A.*) *16th November*, 1944.

598 494.—Waveguide or transmission line composed of materials of selected magnetic and dielectric values chosen to secure an optimum characteristic impedance.

Hazeltine Corporation (*assignees of H. A. Wheeler*). *Convention date* (*U.S.A.*) *16th November*, 1944.

598 640.—Feedback circuit, comprising a phase-control line, an amplitude-control line, and a connecting line, for a valve of the grounded-grid type.

Standard Telephones and Cables Ltd. (assignees of A. G. Kandoian). Convention date (U.S.A.) 28th October, 1944.

599 061.—Frequency-modulator for centimetre waves in which the flexional movements of a piezo-electric crystal are used to vary the tuning capacitance of the master oscillator.

" Patelhold " Patentverwertungs &c. A.G. Convention date (Switzerland) 14th July, 1945.

599 383.—Variable capacitor arranged in series with a high-reactance ' stub ' to serve as a radio-frequency switching device for transmission-line circuits.

Marconi's W.T. Co. Ltd. (assignees of O. McD. Woodward). Convention date (U.S.A.), 27th February, 1943.

SIGNALLING SYSTEMS OF DISTINCTIVE TYPE

598 644.—Keying and control arrangements for multiplex signalling systems using carrier-frequency shift modulation.

F. B. Dehn (communicated by Press Wireless Inc.). Application date 22nd October, 1945.

599 338.—Signalling system in which two intercalated series of pulses are used, and modulation is effected by varying the timing or phasing of one series relatively to the other.

S. Jefferson. Application date 14th September, 1945.

599 339.—Discriminator circuit, based on the selection of odd and even harmonics, for detecting or translating time-modulated pulsed signals.

Standard Telephones and Cables Ltd. (assignees of D. D. Grieg). Convention date (U.S.A.), 11th September, 1944.

600 052.—Receiver for a time-modulated system of pulsed signalling, wherein the detected signals are derived from the movement of the electron beam in a cathode-ray indicator.

Standard Telephones and Cables Ltd. (assignees of E. Labin and D. D. Grieg). Convention date (U.S.A.) 13th November, 1944.

600 148.—Receiving circuit for detecting time-modulated pulsed signals by utilizing an unmodulated carrier frequency which is derived from the incoming signals.

Standard Telephones and Cables Ltd. (assignees of E. Labin and D. D. Grieg). Convention date (U.S.A.) 28th October, 1944.

600 255.—Receiving circuit, including a blocking oscillator, for detecting time-modulated signals which are radiated in pairs of constant-width pulses.

Standard Telephones and Cables Ltd. (assignees of E. Labin and D. D. Grieg). Convention date (U.S.A.) 24th May, 1943.

600 291.—Receiving circuit for detecting time-modulated pulsed signals by comparing them with waves of characteristic form derived from the incoming signals.

Standard Telephones and Cables Ltd. (assignees of D. D. Grieg). Convention date (U.S.A.) 28th September, 1942.

600 742.—Keying device for phase or frequency-modulated waves, and for signalling systems utilizing variable-frequency and variable-width pusles.

Marconi's W.T. Co. Ltd. (assignees of M. G. Crosby). Convention date (U.S.A.) 30th June, 1943.

600 955.—Multiple signalling system in which the different channels carry trains of pulses, modulated by time-displacement, and interleaved.

Standard Telephones and Cables Ltd. (assignees of D. D. Grieg). Convention date (U.S.A.) 19th August, 1944.

601 124.—Detector circuit for pulsed signals which are modulated in amplitude, wherein an electronic switch is arranged to charge and discharge a reactive circuit of given time-constant.

H. R. Whitfield and J. M. Daniels. Application date 5th April, 1945.

601 128.—Receiving system in which a train of time-modulated pulses is first converted into an amplitude-modulated train, and is then rectified.

Standard Telephones and Cables Ltd. (assignees of D. D. Grieg). Convention date (U.S.A.) 29th July, 1944.

601 135.—Modulating system in which a train of pulses is applied to a highly-damped circuit so as to vary the pulse-width in accordance with the applied signal.

Standard Telephones and Cables Ltd. (assignees of D. D. Grieg). Convention date (U.S.A.) 29th July, 1944.

601 442.—Multiplex signalling system for point-to-point communication on frequency-modulated waves of the order of decimetres.

Soc. Francaise Radio-Electrique. Convention date (France) 11th June, 1942.

CONSTRUCTION OF ELECTRONIC-DISCHARGE DEVICES

601 230,—Electrolytic method of depositing a metallic screen and of sealing the joint between the glass bulb and pin-socket of a thermionic valve.

Callender-Suchy Developments Ltd. and K. T. Suchy Application date 21st February, 1945.

601 563.—Construction of a shielded mounting with pin-and-slot connection, for a thermionic valve.

Cinch Manufacturing Corp. (assignees of S. M. del Camp). Convention date (U.S.A.) 20th December, 1944.

SUBSIDIARY APPARATUS AND MATERIALS

598 386.—Direct-coupled relay circuit of the pentode type, designed to respond, without backlash, to an applied voltage difference of given value.

B. M. Hadfield. Application dates 3rd August and 31st December, 1945, and 9th January, 1946.

598 915.—Construction of electronic tube for detecting the presence of radiated energy by its ionizing effects.

Bendix Aviation Corporation. Convention date (U.S.A.) 13th March, 1944.

598 970.—Valve circuit for transmitting voltage variations, between points at different potentials, as in cathode-ray operation, or as a voltage stabilizer.

The General Electric Co. Ltd. and R. B. Mackenzie. Application date 11th May, 1945.

599 080.—Stabilizing the operation of a cross-coupled multivibrator circuit, particularly for frequency control or modulation.

L. E. Currah. Application date 23rd April, 1945.

599 149.—Valve-oscillation generator in which the rate of feedback is automatically controlled so as to maintain a constant output voltage over a wide range of frequencies

The General Electric Co. Ltd., D. W. Berry and R. Mawson. Application date 8th June, 1945.

599 274.—Electro-mechanical resonator formed by a magnetic ring element, which is supported at its nodal points and is energized by a triode oscillator, for which it forms the feedback path.

Marconi's W.T. Co. Ltd. and P. J. Neilson. Application date 15th June, 1945.

WIRELESS ENGINEER

Vol. XXVI APRIL 1949 No. 307

Relatively-moving Charge and Coil

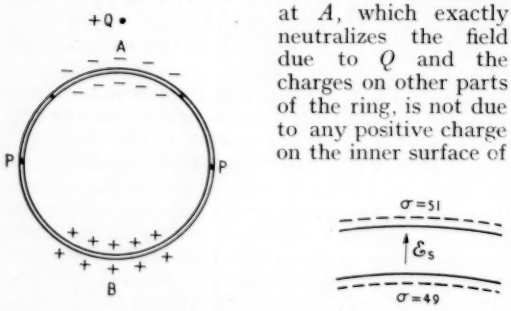
ON p. 140 of this issue we publish a letter from Professor Cullwick criticizing our December Editorial. His criticism appears, however, to be based partly on some misunderstanding. He gives an alleged summary of our argument, but his item (a) is no part of our argument but quite contrary to it. We said that 'all parts of the ring are at the same potential,' but far from maintaining that the fields cancel at all points in the plane of the coil within the loop, we showed in Fig. 1(b) how the electric field enters the ring both on the outside and on the inside, and in Fig. 1(a) the charges are shown on which the field terminates both on the outer and inner surface. Where we differed from Professor Cullwick was with regard to the sign of the charges. He has opposite charges on the outside and inside of the ring and it is fairly obvious that if such a state of affairs were possible the charges would to some extent cancel each other's magnetic effect when in motion. We maintained, and still maintain, that this is quite wrong. He says 'the polarity of the charges on the inner surface is immaterial. This will depend on the geometry of the system, and may be shown by Professor Howe in Fig. 1 of his Editorial or Fig. 2 of my letter in the same issue, according to circumstances.' He had, however, specified the geometry and circumstances as a single loop of thin wire, and for such a loop we maintain that his Fig. 2 was entirely wrong. The polarity of the charges on the inner surface is of very great importance and his failure to realise this is, we believe, the cause of much of his trouble.

With regard to item (b) of his letter, we showed in the September Editorial that, far from cancelling their magnetic effects, the positive and negative induced charges produce within the ring a magnetic field which by a rough approximation we showed to be equal to that produced when the charge was moving and the ring stationary. Although we did not attempt to prove it rigidly we said 'we can safely state that the magnetic field reaches the same maximum value in both cases.' In both cases the observer is regarded as stationary. We realize that this is quite contrary to Professor Cullwick's thesis that the adjacent electric charges are of opposite sign and cancel their magnetic effects, but, as we said above, we regard this as fundamentally wrong.

Under item (c) he correctly states our point of view, viz., that the induced e.m.f. in the ring is the same in both cases. The induced e.m.f. is a physical fact and not like magnetic and electric fields dependent on the motion of the observer. If a charge and a metal ring move relatively to one another, the e.m.f. induced in the ring cannot depend on the idiosyncrasy of any observer. Professor Cullwick admitted this in his letter in our December number but said that difficulties arise in the application of classical theory. Having come to the conclusion that the maximum magnetic flux through the ring is the same as when the ring was stationary and the charge moving, and that a moment later, when the ring has moved away, the flux is reduced to zero, we said 'Would Maxwell have agreed that this magnetic flux could be smuggled somehow in and out of the ring without inducing an e.m.f. in it?' This appears to be the main point of Professor Cullwick's argument, for he maintains that detailed application of Maxwell's theory gives zero e.m.f. We simply applied the formula $e = -d\Phi/dt$ and asked no question as to whether the resultant change of flux was due to the change of flux through a stationary path or to the motion of the ring through the magnetic field. The exact calculation of the two components would be very complicated and we did not consider it necessary. We know that the flux

through the ring is reduced from Φ_{max} to 0, but to what extent this is due to a weakening of the field, and to what extent due to the movement of the ring in the field, we regarded as of no great importance in the present argument.

Professor Cullwick's strange statement that the two components are found to cancel, giving zero e.m.f. indicates that he believes that the magnetic flux can be smuggled out of the ring without inducing an e.m.f. in it, but he gives no suggestion of evidence in support of this statement. If he has calculated the two components and found them equal but of opposite sign, we suspect that one of the signs was wrong ; it would be interesting to know if the components were each equal to $0.5\,d\Phi/dt$; if so, correcting the wrong sign would bring the result into agreement with Maxwell. We doubt, however, if any such detailed calculation has been made. The fundamental error in his Fig. 2 is due to overlooking the fact that the induced charge on the portion of the ring nearer $+Q$ must necessarily be entirely negative while that on the more remote part of the ring, not shown in his diagram, must must be entirely positive. The radial field $\mathcal{E}_s$ within the conductor at A, which exactly neutralizes the field due to Q and the charges on other parts of the ring, is not due to any positive charge on the inner surface of



the ring, but is due entirely to the relatively small difference of surface density of the negative charge on the outer and inner surfaces. This is illustrated in the Figure which shows the charge Q and the thin wire ring. In order to neutralize the induced field at P the induced charges must be distributed roughly as shown. Let us assume that the charge at A corresponds to a surface density of say, 50 units. In order to neutralize the field, in the metal due to the charges, the surface density must be greater on the outer than on the inner surface. We show a portion of the ring at A with surface densities of 51 and 49, but the ratio of the difference to the mean value will depend on the thickness of the ring, and, for the thin wire specified by Professor Cullwick, no great error will be made if one neglects the difference and assumes the same surface density on the outer and inner surface when calculating the magnetic field. It is mainly due to this error that he is led to conclude that ' the problem seems to provide an exception to the accepted principle that Maxwell's theory does not require the assistance of Einstein except when velocities are comparable with the velocity of light.'

In the final paragraph of his letter Professor Cullwick deals with another problem, viz., that of the charge and short-circuited toroid in relative motion, and his criticism in this case is perfectly justified. In our Editorial we said that we could not follow his argument, but his letter now makes the point clear and we agree with him. As the charge moving along the axis approaches the toroid, which we assume to have negligible resistance, the induced current prevents the growth of magnetic flux within the toroid. One may picture two equal and opposite magnetic fluxes within the toroid, one, due to the moving charge, that would occupy the same space if the toroid were not there, and which therefore has no effect on the moving charge, and the other, due to the current, which by its growth induces an electric field tending to accelerate the approaching charge, as Professor Cullwick correctly stated. The energy that would otherwise have been in the magnetic field within the space occupied by the toroid will now be distributed throughout the surrounding space owing to the increased speed of the charge. After passing through the toroid the process will be reversed, but any ohmic losses in the toroid will presumably cause a permanent decrease in the speed of the charge. Our suggestion in the December Editorial that the approaching charge would be retarded was based on a misunderstanding, and the above considerations show that Professor Cullwick was quite right in maintaining that it would be accelerated.

In his previous letter Professor Cullwick mentioned that the ether plays a part in Maxwell's theory, but that nobody believes in it now. This is true, but the part that it played appears to have been taken over by space, which has the same number of letters and, if endowed with the same properties, seems to answer just as well. He also mentioned that many scientists maintain that the idea of a storage of energy in the space around the moving charge is not really tenable. The concepts of the magnetic and electric fields in space with their associated energies have certainly played an important part in the development of electromagnetic theory, and it is difficult to imagine how the subject can be developed without these concepts. Whether they are anything more than concepts is another matter.

G. W. O. H.

GRID CURRENT WITH RC COUPLING

By H. T. Ramsay, M.Sc.(Eng.), M.I.E.E., F.Inst.P.

(Communication from the Research Staff of the M.O. Valve Company at the G.E.C. Research Laboratories, Wembley, England.)

SUMMARY.—A common feature of multivibrator circuits is that the valves forming them are driven into grid current during some part of each cycle. These currents produce, across the coupling capacitors, potentials that play a significant part in the operation and determine the minimum value that the coupling capacitance may be given. The following discussion attempts to analyse these effects in the general case of a valve driven into grid current by a square-wave through capacitance coupling.

THE coupling elements in Fig. 1 are linear and therefore, if the valves are operated over a reasonably linear part of their characteristics, it is a simple matter to calculate the output from the input waveform; similarly there is no difficulty in calculating the specifications of C, R, and r.

If, however, the second grid is driven into grid current, a non-linear element is introduced and the calculations become less obvious. This situation is frequently met, a good example being the circuit shown in Fig. 2. This circuit is based on the Schmitt trigger[1], and is required to produce square waves with very sharp fronts from a sine wave input, the frequency of which may be varied over a wide range. In this circuit, the voltage on the first anode has the form shown in Fig. 3; i.e., it is roughly a square wave. This voltage drives the second grid alternately to cut-off and then into grid current, thus producing the square-wave output. The merit of the circuit lies, of course, in the cathode coupling which results in extremely sharp wavefronts in the output.

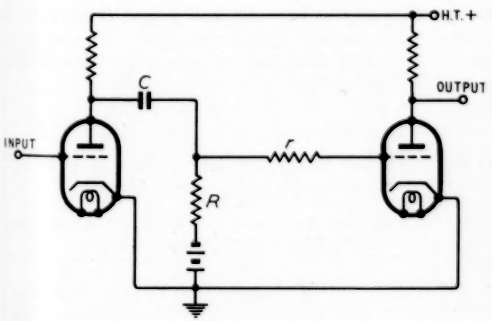


Fig. 1. *The elementary circuit.*

During the intervals 0–T, $2T$–$3T$, etc. (see Fig. 3), the grid drive current builds up a charge on C, so producing a potential difference which opposes the driving e.m.f.; consequently the drive current falls steadily until the anode voltage falls sharply at T, $3T$, etc., by the operation of the circuit. If the frequency is reduced, a value is ultimately reached at which the charge built up on C produces a potential great enough to stop the further flow of grid current against the cathode potential. A further reduction of frequency then produces a voltage on the driven grid as shown in Fig. 4(b) and the output is correspondingly distorted. The frequency minimum so reached is evidently a function of C, R, and r among other quantities and it has been thought worthwhile to investigate this relation as there seemed to be no other convenient way of finding the best values for these components.

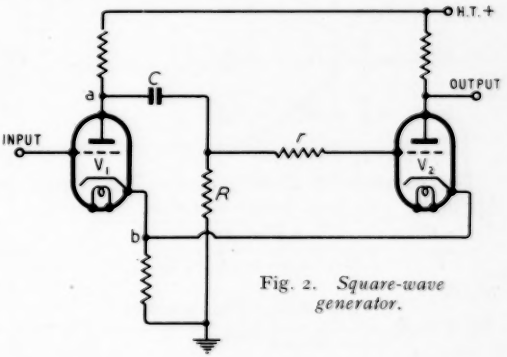


Fig. 2. *Square-wave generator.*

As far as currents in the coupling elements are concerned, Fig. 5 is an equivalent of Fig. 2 provided that the voltage sources $V(t)$ and $E(t)$ produce the same potentials to earth as those at points a and b of Fig. 2. It is assumed therefore that $V(t)$ and $E(t)$ have the waveforms shown in Fig. 10(a) and (b) respectively when feeding into the load provided by the circuit of Fig. 5. The analysis may be simplified, without serious loss of equivalence, by assuming $V(t)$ and $E(t)$ to have the waveforms shown in Fig. 6.

MS accepted by the Editor, January 1948

The physical processes that are to be investigated are as follows. In the intervals 0–T, $2T$–$3T$, etc., the potential on the anode of V_1 charges C through the resistances R and r in parallel. In the intervals T–$2T$, etc., C discharges through R alone, because the rectifying action of the grid prevents current from flowing through r. But the mean current through C must be zero when the circuit is in steady operation, accordingly a steady charge is built up on C which produces a potential so biasing the voltage waveform that the mean current through C in each half cycle is the same, despite the unbalance between the charge and the discharge resistances.

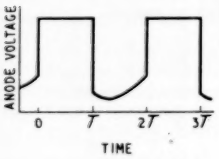


Fig. 3. *Waveform of voltage on driving anode.*

Applying Kirchhoff's Laws to the arrangement of Fig. 5 gives

$$V(t) = \frac{q}{C} + ri_2 + E(t) \quad \ldots \quad (1)$$

$$V(t) = \frac{q}{C} + R(i_1 - i_2) \quad \ldots \quad (2)$$

$$dq/dt = i_1 \quad \ldots \quad (3)$$

The solution of these equations is shown in Appendix 1 to lead to the equation

$$\frac{nE}{V} = \frac{\beta . F(\alpha, \beta)}{\beta - \alpha + \alpha . F(\alpha, \beta)} \quad \ldots \quad (18)$$

where

$$F(\alpha,\beta) = \frac{e^{-\beta/\alpha(\beta-\alpha)} - e^{-(2\beta-\alpha)/\alpha(\beta-\alpha)}}{1 - e^{-(2\beta-\alpha)/\alpha(\beta-\alpha)}}$$

$$\alpha = CR/T$$

and

$$\beta = \frac{CR}{T} + \frac{Cr}{T}$$

T is the duration of the half period of the frequency at which defective operation (as described above) appears; i.e., of the lower-frequency limit of the unit.

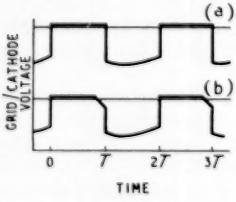


Fig. 4. *Voltage on second grid above (a) and below (b) the minimum frequency.*

Equation 18 states the relation between α, β, and nE/V for the diode current just to fall to zero at the end of the half-period T. Alternatively, if a stage developing a rectangular signal of amplitude V is required to drive the following valve into grid current for a full half-period against a bias nE, Equ. 18 gives the pairs of values of α and β that just suffice. The curves resulting from the evaluation of this equation are shown in Fig. 7; none of those plotted rises above 0.46 and it may be shown that Equ. 18 has a limiting value of 0.5, which it approaches when α, β, and $(\beta-\alpha)$ are very great. The form of the curves shows that, for a particular value of nE/V, there is minimum value of β that may be employed; this value being that which produces a curve with a maximum equal to the required value of nE/V. The values of α and β corresponding to the maxima in Fig. 7 are given in Table I; these values are obtained by difference methods from the tabular values used in plotting the curves and have a maximum error of 0.0005. The table also gives the corresponding values of Cr/T.

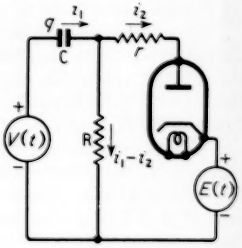


Fig. 5. *The equivalent arrangement.*

It should be noted that Fig. 7 shows that values of α and β may be chosen to meet any value of nE/V less than 0.5. If nE/V is 0.5 or greater, it is not possible to arrange that grid current persists for a full half-cycle.

Now let it be supposed that the coupling elements R, r and C of a particular equipment are to be investigated and that, as is usual, it is desired to reduce C to a minimum. T and nE/V will be part of the prescribed data and therefore Equ. 18 may be written

$$V/nE - 1 = \text{a constant} = f(\alpha, \beta) \quad \ldots \quad (18a)$$

Given that a pair of values of α and β have been chosen that satisfy Equ. 18a, the values of R, r, and C may be changed in any way that leaves α and β unaltered.

Since

$$\alpha = CR/T \text{ and } \beta = C(r + R)/T$$

it is clear that whatever the values of α and β,

TABLE I

Pairs of values of Cr/T and CR/T giving a minimum value of β for a given value of nE/V.

nE/V	β	α (CR/T)	$\beta-\alpha$ (Cr/T)
0.032	1	0.547	0.453
0.169	2	1.124	0.876
0.269	3	1.707	1.293
0.318	4	2.288	1.712
0.377	6	3.461	2.539
0.408	8	4.641	3.359
0.426	10	5.793	4.207

the smallest value of C is obtained by using the greatest value of $(R + r)$. However $(R + r)$ is the external grid-cathode resistance, a quantity for which valve makers invariably specify a maximum. Thus the value of $(R + r)$ is determined, it remains to divide the available resistance between R and r. But it has already been shown that the smallest value of β, and hence of C since $(R + r)$ is already fixed, is permitted when α lies on the curve defined by the values in Table I. Reverting to Fig. 7 the flatness of the curves in the neighbourhood of the maxima invites practical approximations, an obvious one being to take the maxima as given by values of Cr/T equal to those given by Table I for CR/T. This slightly increases β above the true minimum but

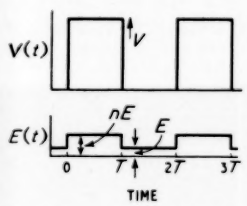


Fig. 6. *Waveforms in the equivalent circuit.*

TABLE II

Minimum values of CR/T for given values of nE/V when R is made equal to r.

nE/V	0.0313	0.1652	0.2585	0.3141	0.3498
CR/T	0.5	1.0	1.5	2.0	2.5
nE/V	0.3743	0.3921	0.4054	0.4160	0.4244
CR/T	3.0	3.5	4.0	4.5	5.0

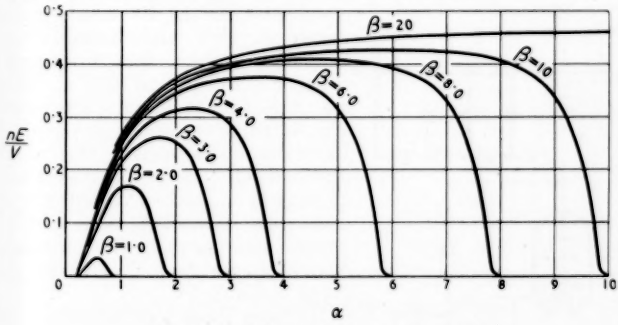


Fig. 7. *n/EV as a function of α and β.*

makes R equal to r. The appropriate values are given in Table II and plotted in Fig. 8. It follows that, for all practical purposes, the smallest value of C is obtained by making both R and r equal to half the maximum permitted grid-cathode resistance.

It is sometimes necessary to restrict the value of either R or r. For instance, it may be necessary to maintain at a low value the time constant formed by r and the input capacitance of the driven grid. A question then arises as to the value of R that permits the smallest value for C when r is restricted and the two remaining variables in Equ. 18 (i.e., T and nE/V) have some assigned value. Evidently T and nE/V become, for present purposes, any positive constants subject only to the restriction that nE/V must be less than 0.5. Equ. 18(a) may be applied to these conditions, and the problem becomes one of investigating the way in which C changes in this equation when r is held constant and R allowed to vary. This investigation is made in Appendix II,

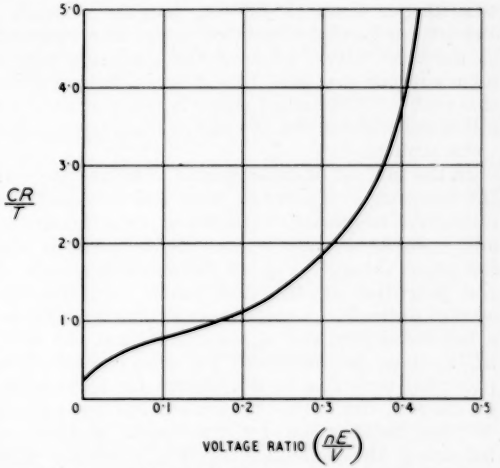


Fig. 8. *Minimum values of CR/T for given values of nE/V; R is made equal to r.*

where it is shown that, whatever the value assigned to r, the smallest C results from increasing R to the maximum possible extent. Similarly if R is assigned a value, the smallest C results from increasing r to the maximum extent. In other words, C is minimized by arranging R and r to be approximately equal and their sum to be as great as possible: if it is not possible to make R and r approximately equal, the smallest C is still obtained by making their sum as great as possible.

Experimental Verification

The above calculations were checked by experiment on a small pulse generator. The circuit (shown in Fig. 9) is an elaboration of the arrangement of Fig. 2. A diode has been inserted

in the grid circuit of the second valve, so that the voltage across the cathode resistance of this valve appears as the minimum grid bias. The mean anode current may then be set to the rated maximum for the valve type by adjusting the value of this resistance which, incidentally, is too low to require a capacitive by-pass. A further change is that the anode load of this valve has been modified by adding an artificial line. Then, to a sufficiently rapid disturbance, the anode load presents the parallel resistance of R_0 and Z_0 for twice the delay time of the line and then collapses to approximately zero. Provided, therefore, that the grid signal persists longer than this time, the duration of the output impulse is controlled by the artificial line, and may be varied by manually switching-in lines of differing lengths. Finally, a short line has been inserted in the anode circuit of the first valve, because the initiating signal for a cathode-ray tube time base is derived from this point and the short delay between it and the following grid enables the rise of the output pulse to be studied.

In the first set of experiments, various values of the coupling components were inserted and the minimum frequency determined experimentally and compared with values deduced from the foregoing theory. Fig. 10 shows oscillograms of the potential on the first anode, and on the second cathode. Comparison with Fig. 6 shows a fair similarity and it is evident that the ratio nE/V may be measured on the traces. The potential zero line is not shown on the oscillograms and the actual measurements were made on the oscilloscope by the usual method of balancing the deflection against a steady shift voltage applied to the opposite oscilloscope plate. The minimum frequency was determined by watching the grid-cathode potential of the second valve while running down the input frequency, so finding the frequency at which this potential changes from the form of Fig. 4(a) to that of (4b). Fig. 11 shows oscillograms of the grid-cathode potential above and below the minimum frequency. The results of these experiments are given in Table III.

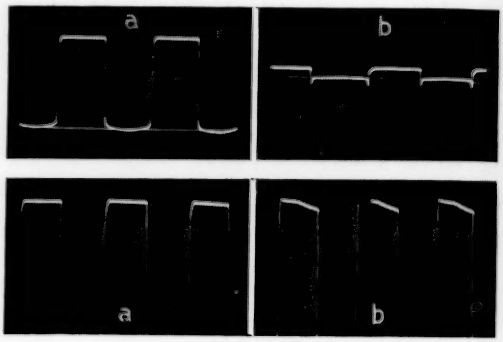


Fig. 10 (above). *Potential on the first anode (a); on the second anode (b).*

Fig. 11 (below). *Grid-cathode potential at frequencies above (a) and below (b) the minimum.*

The discrepancies between the measured and the calculated values seem reasonable enough, allowing for the differences between the waveforms of Figs. 6 and 10.

Experiments were then made to check the conclusion that, to meet given values of nE/V and the minimum frequency when the total grid-cathode resistance $(R + r)$ is fixed, r and R should equal each other if C is to be as small as possible. In these experiments the total grid-cathode resistance was fixed at 500 kΩ and r then varied over a wide range; at each value of r, C was adjusted to make the minimum frequency 300 c/s. The results are given in Table IV.

If, from these results, $(R - r)$ is plotted against C, it will be seen that, to the accuracy of the experiments, the minimum capacitance occurs when $(R - r)$ is zero.

The final experiments were made to confirm that, with given values of nE/V and the minimum frequency and when r is at some fixed value C is minimized by making R as

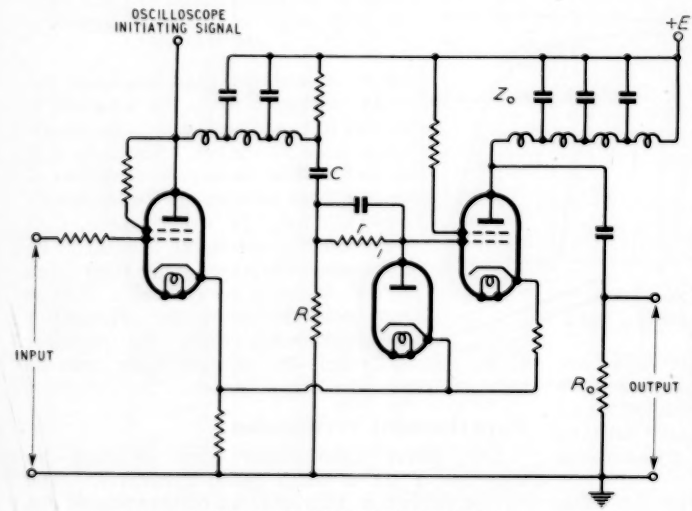


Fig. 9. *Small impulse generator.*

large as other considerations permit. In these experiments, r was fixed at 30 kΩ and R then varied over a wide range; at each value of R, C was adjusted to make the minimum frequency 300 c/s. Table V gives the results.

TABLE III

Comparison of experimentally and theoretically determined values of the minimum frequency, with R and r each of 50 kΩ.

C (μF)	nE/V	Minimum frequency (c/s)	
		Experimental	Theoretical
0.01	0.069	680	645
0.02	0.073	300	330
0.10	0.073	70	65.9

TABLE IV

Optimum division of the total grid-cathode resistance (500 kΩ) between R and r.

Minimum frequency (c/s)	r (kΩ)	R (kΩ)	C (μF)
300	30	470	0.018
300	50	450	0.010
300	100	400	0.006
300	200	300	0.0055
300	300	200	0.0055
300	400	100	0.009
300	450	50	0.020
300	470	30	0.030

TABLE V

Optimum value for R when r is fixed; minimum frequency = 300 c/s, r = 30 kΩ.

R (Ω)	C (μF)
30,000	0.030
50,000	0.025
100,000	0.020
470,000	0.018

These results evidently support the theoretical result.

Acknowledgment

In conclusion, the author desires to tender his acknowledgment to the M.O. Valve Co. Ltd., on whose behalf the work described in this publication was carried out.

REFERENCE

[1] "A Thermionic Trigger," by O. H. Schmitt. *Journal of Scientific Instruments*, 1938, Vol. 15, p. 24.

APPENDIX I

The solution of the circuit equations.

The equations to be solved are

$$V(t) = \frac{q}{c} + ri_2 + E(t) \quad .. \quad .. \quad .. \quad .. \quad (1)$$

$$V(t) = \frac{q}{c} + R(i_1 - i_2) \quad .. \quad .. \quad .. \quad .. \quad (2)$$

$$\frac{dq}{dt} = i_1 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (3)$$

The Laplace transforms are therefore

$$[V] - [E] = + r[i_2] + \frac{[q]}{C}$$

$$[V] = R[i_1] - R[i_2] + \frac{[q]}{C}$$

$$- q^{(0)} = [i_1] \qquad - p[q]$$

where $[V]$ is the Laplace transform of $V(t)$, $[i_1]$ is the transform of i_1, and so on. Also $q^{(0)}$ is the value of q at the time 0, $q^{(T)}$ is the value of q at the time T, and similarly for the other variables.

It follows from the last three equations that

$$[i_1] = \frac{pC[V] - q^{(0)}}{1 + pCR_p} - \frac{pC[E]}{r} \cdot \frac{R_p}{1 + pCR_p} \quad .. \quad (4)$$

$$[i_2] = \frac{pC[V] - q^{(0)}}{r} \cdot \frac{R_p}{1 + pCR_p} - \frac{[E]}{r + R} \cdot \frac{1 + pCR}{1 + pCR_p} \quad .. \quad .. \quad .. \quad (5)$$

$$[q] = \frac{C[V]}{1 + pCR_p} + \frac{q^{(0)} CR_p}{1 + pCR_p} - \frac{C[E]}{r} \cdot \frac{R_p}{1 + pCR_p} \quad .. \quad .. \quad .. \quad (6)$$

Where

$$R_p = \frac{rR}{r + R}$$

Inspection of Fig. 6 shows that

$V(t) = V$ for $0 < t < T$

$V(t) = 0$ for $T < t < 2T$

$E(t) = nE$ for $0 < t < T$, where n is some positive constant.

and $E(t) = E$ for $T < t < 2T$.

It follows that, in the interval $0 < t < T$,

$$[V] = V/p, \text{ and } [E] = nE/p.$$

Inserting these results in Equs. (4) and (5), and inverting the transforms gives (for the interval $0 < t < T$)

$$i_1 = \frac{V - q^{(0)} C}{R_p} \cdot e^{-t/CR_p} - \frac{nE}{r} \cdot e^{-t/CR_p} \quad .. \quad (8)$$

$$i_2 = \frac{(V - q^{(0)}/C - nE)}{r} \cdot e^{-t/CR_p} - \frac{nE}{r + R} \cdot (1 - e^{-t/CR_p}) \quad .. \quad .. \quad .. \quad (9)$$

$$\frac{q}{C} = V - \frac{nER}{r + R} - \left(V - \frac{nER}{r + R} - \frac{q^{(0)}}{C}\right) \cdot e^{-t/CR_p} \quad (10)$$

From Equ. 10, it is seen that the change in q during the interval $0 < t < T$ equals

$$q^{(T)} - q^{(0)} = C\left(V - nE \cdot \frac{R_p}{r} - \frac{q^{(0)}}{C}\right)(1 - e^{-T/CR_p}) \; ..$$

whence

$$\frac{q^{(T)}}{c} = \left(V - nE\frac{R_p}{r}\right)(1 - e^{-T/CR_p}) + \frac{q^{(0)}}{C} \cdot e^{-T/CR_p} \quad .. \quad .. \quad .. \quad (11)$$

Before proceeding to the next interval, it is worth noting that if i_2 falls to zero at t_0, Equs. (8), (9), (10), and (11) are invalid for values of T greater than t_0. This is because the operation of the diode (see Fig. 5) isolates the branch containing r at this instant. How-

WIRELESS PATENTS

A Summary of Recently Accepted Specifications

The following abstracts are prepared, with the permission of the Controller of H.M. Stationery Office, from Specifications obtainable at the Patent Office, 25, Southampton Buildings, London, W.C.2, price 2/- each.

AERIALS AND AERIAL SYSTEMS

599 387.—Trailing aerial for aircraft in which the tuning is automatically controlled through the winding and unwinding gear.

Autophon A.G. Convention date (Switzerland) 24th December, 1943.

DIRECTIONAL AND NAVIGATIONAL SYSTEMS

598 609.—Pulsed signalling system for visually indicating to the pilot of a mobile craft his position, in terms of azimuth and distance, relatively to a fixed reference point.

Hazeltine Corporation (assignees of A. V. Loughren). Convention date (U.S.A.) 15th September, 1944.

598 672.—Automatic gain-control system, particularly for a radiolocation receiver in which the echo signals are presented side-by-side, or on separate indicator tubes, for comparison.

J. V. Dunworth. Application date 28th June, 1945.

598 701.—Radiolocation equipment for centimetre waves, wherein the time interval is accurately measured in terms of the number of complete cycles of an auxiliary synchronized oscillatory voltage.

J. D. Cockcroft, P. E. Pollard, B. Newsom, The British Thomson-Houston Co. Ltd. and G. S. C Lucas. Application date 30th March, 1943.

598 900.—Radio signalling system for marking-out a water-channel, or other tortuous course, by the radiation from a number of electromagnetic 'horn' aerials placed at strategic points.

Western Electric Co. Inc. Convention date (U.S.A.) 5th April, 1940.

598 948.—Radiolocation receiver wherein automatic gain-control is applied to increase gradually the sensitivity of the indicator to the more-distant objects.

Sperry Gyroscopic Co. Inc. (assignees of G. E. White). Convention date (U.S.A.) 22nd January, 1943.

598 973.—Circuit arrangement for generating the exponential voltages used for controlling the cathode-ray indicators in radiolocation.

Marconi's W.T. Co. Ltd. (assignees of M. H. Mesner). Convention date (U.S.A.) 20th August, 1942.

598 977.—Beam aerial comprising a given length of waveguide with spaced and graded apertures, which are backed by rotary reflecting discs in order to swing the beam to-and-fro.

Marconi's W.T. Co. Ltd. (assignees of H. B. de Vore). Convention date (U.S.A.) 9th June 1944.

599 087.—Direction-finding system, operating on pulsed signals and comprising a switch-controlled receiver for separately comparing the response from two aerials of different characteristics.

B. S. Smith and F. E. J. Girling. Application date 30th June, 1945.

599 114.—Radiolocation apparatus utilizing the Doppler frequency-shift, wherein a .local reference-frequency having a fixed phase-relation with the exploring pulses is derived from supersonic oscillations set up by a piezo-electric crystal along a mechanical line resonator.

A. E. Solley. Application date 13th September, 1945.

599 389.—Automatic gain-control system applied to the echo signals received in the different sectors of a rotary exploring beam, in radiolocation.

J. M. Robson. Application date 2nd May, 1945.

599 497.—Coupling bridge-network, including a phasing device, arranged to allow two aerials having different polar diagrams to be continuously interchanged, say for radiolocation.

C. H. Westcott. Application date 9th January, 1945.

599 551.—Direction finder using a rotating aerial and a visual indicator, wherein auxiliary optical means are provided for adjusting the normal alignment of the signal on the indicator.

Standard Telephones and Cables Ltd. (assignees of G. S. Burroughs). Convention date (U.S.A.) 26th January, 1944.

600 166.—Automatic gain-control system, particularly for regulating the output of a radiolocation receiver.

Sperry Gyroscope Co. Inc. Convention date (U.S.A.), 29th September, 1944.

600 327.—Directive aerial system comprising a parabolic reflector with a focal length only one-fifth that of the carrier wave, and a broad-band radiating element located at the focal point.

Marconi's W.T. Co. Ltd. (assignees of G. H. Brown and O. McD. Woodward Junr.). Convention date (U.S.A.) 22nd January, 1944.

600 433.—Aerial for ultra-short waves in which the action of an electromagnetic horn or parabolic mirror is controlled or modified by resonant slots.

H. G. Booker. Application date 31st October, 1945.

600 483.—Direction finder with a switched c.r. indicator which is fed through a common amplifying channel, wherein periodic damping is applied in order to prevent distortion.

F. Caplin and J. H. Bagley. Application date 7th June 1945.

600 494.—Radiolocation equipment in which the gain control is automatically varied between given limits by a multivibrator device, in order to minimize the effect of ground signals, 'clutter,' and the like.

R. H. A. Carter. Application date 14th September, 1945.

600 512.—Control circuit for the electrodes of a c.r. indicator tube, particularly for preventing misalignment between the line-of-sight and the beam axis, in radiolocation.

R. S. Webley. Application date 2nd May, 1945.

600 514.—Time base and marker pulse generating and triggering circuits, applied to the measurement of short time intervals, particularly in radiolocation.

A. Harrison. Application date 2nd June, 1945.

600 546.—Pulsed radiolocation equipment for use on a mobile craft, wherein provision is made to offset the effects of yawing and pitching.

Sperry Gyroscope Co. Inc. (*assignees of J. J. Caldwell, Jr. and R. F. Hays, Jr.*). *Convention date* (*U.S.A.*) *19th May*, 1942.

600 586.—Radiolocation equipment in which the echo signals from two flat exploring beams are applied to follow the course, and indicate the origin, of a shell or other projectile.

A. E. Kempton and D. R. Chick. Application date 23rd October, 1945.

600 622.—Direction finder in which signals from different aerials are fed to a cathode-ray indicator through separate amplifiying channels, the gain of which is automatically equalized by means of an auxiliary oscillator.

Soc. Francaise Radio-Electrique. Convention date (*France*), 14th February, 1942.

600 633.—Adapting a normally non-directive aerial of the ' Alford Loop ' type so that its radiation pattern is made directive.

Standard Telephones and Cables Ltd. (*assignees of A. G. Kandoian*). *Convention date* (*U.S..A.*) *17th July*, 1944,.

RECEIVING CIRCUITS AND APPARATUS

(*See also under Television*)

598 398.—Super-regenerative receiver, with cathode load resistance, for responding to interrogating signals and triggering an identification signal in reply.

Ferranti Ltd., M. K. Taylor and F. C. Williams. Application date 5th November, 1945.

598 399.—Coupling arrangement between the super-regenerative receiver and the triggered transmitter of an automatic ' responder ' or identification unit.

Ferranti Ltd. and M. K. Taylor. Application date 5th November, 1945.

598 400.—Control arrangement for stabilizing the operation of an identification unit designed to respond automatically to interrogating signals.

Ferranti Ltd., F. C. Williams and J. R. Whitehead. Application date 5th November, 1945.

598 401.—Controlling the sensitivity of the super-regenerative receiver, and the stability of the associated transmitter, in an identification unit which responds automatically to interrogating signals.

Ferranti Ltd. and H. Wood. Application date 5th November, 1945.

598 662.—Selective coupling arrangement, say for a superheterodyne, in which an increase in coupling causes the separate peaks of the respective circuits to move further apart.

Murphy Radio Ltd., P. C. Cullen and C. C. Holmes. Application date 24th November, 1944.

598 721.—Band-pass coupling, comprising first and second transformers, wherein the secondary winding of the first is reactively coupled to the primary winding of the second, and conversely, in order to widen the response curve.

K. W. Horner. Application date 7th September, 1945.

598 807.—Super-regenerative receiver wherein the input circuit is coupled to the oscillatory circuit of the amplifier in rhythm with the quenching frequency, in order to prevent re-radiation.

" *Patelhold" Patentverwertungs &c. A.G. Convention date* (*Switzerland*), *11th September*, 1944.

598 868.—Receiving circuit for frequency-modulated signals, comprising a cathode-follower amplifier and a pair of diodes, arranged so as to throw very little load across the input circuit.

Marconi's W.T. Co. Ltd. (*assignees of K. Schlesinger*). *Convention date* (*U.S.A.*), *21st September*, 1944.

599 201.—Intervalve coupling for h.f. and i.f. amplifiers in which a tapped inductance is used both to provide a band-pass characteristic and to match the valve impedances.

P.R.T. Laboratories Ltd. and W. E. Benham. Application dates 11th May, 1945, *and 11th April*, 1946.

599 341.—Construction, mounting, and processing of contact detectors of the silicon type, particularly for high-frequency work.

Western Electric Co. Inc. Convention date (*U.S.A.*) *24th March*, 1944.

599 424.—Super-regenerative receiver in which a grounded-grid valve is directly coupled to the regenerative circuits and serves as a buffer to prevent re-radiation from the aerial.

Hazeltine Corporation (*assignees of B. D. Loughlin*). *Convention date* (*U.S.A.*) *1st March*, 1945.

TELEVISION CIRCUITS AND APPARATUS

FOR TRANSMISSION AND RECEPTION

598 364.—Projecting large-scale television pictures from a photographic film that is prepared from the image produced on the sensitive screen of a cathode-ray tube.

F. B. Dehn (*communicated by A. B. Dumont Laboratories Inc.*). *Application date 13th March*, 1945.

598 792.—Television receiver in which the fluorescent screen of a cathode-ray tube is of globular form, and is rotated synchronously with the scanning speed, in order to project the picture on to a viewing screen.

The General Electric Co. Ltd. and L. C. Jesty. Application date 10th September, 1945.

598 802.—Television receiver in which the electron stream scans a thin film of oil on a slowly-rotating glass disc, so as to vary its transparency to an external source of light, thus allowing the picture to be projected on to an external screen.

W. Ehrenberg. Application date 11th September, 1945.

599 134.—Colour television system in which the effective areas of the different filters are automatically interrelated so as to preserve a correct colour balance for all adjustments.

Radio Corporation of America. Convention date (*U.S.A.*) *29th November*, 1943.

TRANSMITTING CIRCUITS AND APPARATUS

(*See also under Television*)

598 375.—Coupling-device between two offset and relatively-rotatable waveguides, in which the opposite ends of a complex coaxial-line connector serve as probes.

R. G. Garfitt. Application date 7th May, 1945.

598 493.—Waveguide or transmission line composed of materials having graded values of magnetic permeability and dielectric constant, chosen to secure a predetermined mode of propagation.

Hazeltine Corporation (*assignees of H. A. Wheeler*). *Convention date* (*U.S.A.*) *16th November*, 1944.

598 494.—Waveguide or transmission line composed of materials of selected magnetic and dielectric values chosen to secure an optimum characteristic impedance.

Hazeltine Corporation (*assignees of H. A. Wheeler*). *Convention date* (*U.S.A.*) *16th November*, 1944.

598 640.—Feedback circuit, comprising a phase-control line, an amplitude-control line, and a connecting line, for a valve of the grounded-grid type.

Standard Telephones and Cables Ltd. (assignees of A. G. Kandoian). Convention date (U.S.A.) 28th October, 1944.

599 061.—Frequency-modulator for centimetre waves in which the flexional movements of a piezo-electric crystal are used to vary the tuning capacitance of the master oscillator.

" Patelhold " Patentverwertungs &c. A.G. Convention date (Switzerland) 14th July, 1945.

599 383.—Variable capacitor arranged in series with a high-reactance ' stub ' to serve as a radio-frequency switching device for transmission-line circuits.

Marconi's W.T. Co. Ltd. (assignees of O. McD. Woodward). Convention date (U.S.A.), 27th February, 1943.

SIGNALLING SYSTEMS OF DISTINCTIVE TYPE

598 644.—Keying and control arrangements for multiplex signalling systems using carrier-frequency shift modulation.

F. B. Dehn (communicated by Press Wireless Inc.). Application date 22nd October, 1945.

599 338.—Signalling system in which two intercalated series of pulses are used, and modulation is effected by varying the timing or phasing of one series relatively to the other.

S. Jefferson. Application date 14th September, 1945.

599 339.—Discriminator circuit, based on the selection of odd and even harmonics, for detecting or translating time-modulated pulsed signals.

Standard Telephones and Cables Ltd. (assignees of D. D. Grieg). Convention date (U.S.A.), 11th September, 1944.

600 052.—Receiver for a time-modulated system of pulsed signalling, wherein the detected signals are derived from the movement of the electron beam in a cathode-ray indicator.

Standard Telephones and Cables Ltd. (assignees of E. Labin and D. D. Grieg). Convention date (U.S.A.) 13th November, 1944.

600 148.—Receiving circuit for detecting time-modulated pulsed signals by utilizing an unmodulated carrier frequency which is derived from the incoming signals.

Standard Telephones and Cables Ltd. (assignees of E. Labin and D. D. Grieg). Convention date (U.S.A.) 28th October, 1944.

600 255.—Receiving circuit, including a blocking oscillator, for detecting time-modulated signals which are radiated in pairs of constant-width pulses.

Standard Telephones and Cables Ltd. (assignees of E. Labin and D. D. Grieg). Convention date (U.S.A.) 24th May, 1943.

600 291.—Receiving circuit for detecting time-modulated pulsed signals by comparing them with waves of characteristic form derived from the incoming signals.

Standard Telephones and Cables Ltd. (assignees of D. D. Grieg). Convention date (U.S.A.) 28th September, 1942.

600 742.—Keying device for phase or frequency-modulated waves, and for signalling systems utilizing variable-frequency and variable-width pusles.

Marconi's W.T. Co. Ltd. (assignees of M. G. Crosby). Convention date (U.S.A.) 30th June, 1943.

600 955.—Multiple signalling system in which the different channels carry trains of pulses, modulated by time-displacement, and interleaved.

Standard Telephones and Cables Ltd. (assignees of D. D. Grieg). Convention date (U.S.A.) 19th August, 1944.

601 124.—Detector circuit for pulsed signals which are modulated in amplitude, wherein an electronic switch is arranged to charge and discharge a reactive circuit of given time-constant.

H. R. Whitfield and J. M. Daniels. Application date 5th April, 1945.

601 128.—Receiving system in which a train of time-modulated pulses is first converted into an amplitude-modulated train, and is then rectified.

Standard Telephones and Cables Ltd. (assignees of D. D. Grieg). Convention date (U.S.A.) 29th July, 1944.

601 135.—Modulating system in which a train of pulses is applied to a highly-damped circuit so as to vary the pulse-width in accordance with the applied signal.

Standard Telephones and Cables Ltd. (assignees of D. D. Grieg). Convention date (U.S.A.) 29th July, 1944.

601 442.—Multiplex signalling system for point-to-point communication on frequency-modulated waves of the order of decimetres.

Soc. Francaise Radio-Electrique. Convention date (France) 11th June, 1942.

CONSTRUCTION OF ELECTRONIC-DISCHARGE DEVICES

601 230,—Electrolytic method of depositing a metallic screen and of sealing the joint between the glass bulb and pin-socket of a thermionic valve.

Callender-Suchy Developments Ltd. and K. T. Suchy Application date 21st February, 1945.

601 563.—Construction of a shielded mounting with pin-and-slot connection, for a thermionic valve.

Cinch Manufacturing Corp. (assignees of S. M. del Camp). Convention date (U.S.A.) 20th December, 1944.

SUBSIDIARY APPARATUS AND MATERIALS

598 386.—Direct-coupled relay circuit of the pentode type, designed to respond, without backlash, to an applied voltage difference of given value.

B. M. Hadfield. Application dates 3rd August and 31st December, 1945, and 9th January, 1946.

598 915.—Construction of electronic tube for detecting the presence of radiated energy by its ionizing effects.

Bendix Aviation Corporation. Convention date (U.S.A.) 13th March, 1944.

598 970.—Valve circuit for transmitting voltage variations, between points at different potentials, as in cathode-ray operation, or as a voltage stabilizer.

The General Electric Co. Ltd. and R. B. Mackenzie. Application date 11th May, 1945.

599 080.—Stabilizing the operation of a cross-coupled multivibrator circuit, particularly for frequency control or modulation.

L. E. Currah. Application date 23rd April, 1945.

599 149.—Valve-oscillation generator in which the rate of feedback is automatically controlled so as to maintain a constant output voltage over a wide range of frequencies

The General Electric Co. Ltd., D. W. Berry and R. Mawson. Application date 8th June, 1945.

599 274.—Electro-mechanical resonator formed by a magnetic ring element, which is supported at its nodal points and is energized by a triode oscillator, for which it forms the feedback path.

Marconi's W.T. Co. Ltd. and P. J. Neilson. Application date 15th June, 1945.